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**(54) Method of signal analysis employing histograms**

(57) A method of signal analysis for establishing display scaling and stable triggering for digital oscilloscopes employs amplitude and time histograms to analyze an input signal and establish the required conditions. An amplitude histogram analysis method is first used to determine the minimum and maximum amplitude levels of the signal, and then a time histogram analysis method is used to determine the predominant time intervals occurring in the input signal. From these two analyses, vertical gain adjustments and time base settings adjustments may be made either manually or automatically, and triggering levels established to provide a stable, usable display.

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## Description

### Background of the Invention

This invention relates generally to scaling and triggering arrangements for oscilloscopes, and in particular to a system for analyzing input signals using amplitude and time histograms to establish vertical sensitivity, time base setting, and triggering levels for oscilloscopes to provide stable, viewable displays.

Typical digital oscilloscopes rely on the user to manually adjust vertical and horizontal scale factors, and to adjust the triggering level to achieve stable triggering. More sophisticated oscilloscopes utilize digital comparators to detect peak amplitude values of input signals automatically to establish vertical sensitivity and level-crossing trigger detection automatically to establish a time base setting as well as a triggered display.

Achieving stable triggering has always been one of the more difficult tasks in operating an oscilloscope. In analog oscilloscopes, a trigger comparator issues trigger signals to start an associated time base sweep when an input signal matches a voltage level adjusted arbitrarily by a user, and often the precise triggering point is difficult to ascertain. In digital oscilloscopes, issuance of a valid trigger signal often is used to stop digital acquisition of an input signal in a so-called post-trigger mode, freezing the contents of an associated waveform memory so that a later reconstructed waveform display will show the user events leading up to the triggering event. Thus, establishing a set of conditions upon which a valid trigger will be issued can be complicated, even for experienced users. Accordingly, many attempts have been made to provide automatic triggering to relieve the user of this often frustrating task.

Many users are only interested in a stable image of an input signal to see whether or not something is wrong with the shape of the signal. For these users, operating the instrument is only a bothersome necessity. As oscilloscopes are being simplified and designed for use by lesser experienced personnel, a simple and inexpensive method of establishing automatic vertical sensitivity, time base setting, and triggering is desired.

### Summary of the Invention

In accordance with the present invention, a method of achieving automatic signal scaling and stable triggering for digital storage oscilloscopes employs amplitude and time histograms to analyze an incoming signal.

An untriggered, or free-running, incoming signal is converted to digital representations, or samples, by an analog-to-digital converter (ADC) in the conventional manner and processed by an acquisition processor. The digitized samples are first analyzed to form an amplitude histogram by counting the number of occurrences of each discrete amplitude value. The minimum (MIN) and maximum (MAX) signal excursions may be quickly determined by this method. The MIN and MAX amplitude val-

ues may then be used to adjust the vertical sensitivity of the input amplifier to result in an on-screen displayed waveform of an appropriately-scaled (on the vertical axis) reconstructed signal. Also, lower and upper triggering levels may be established within the peak-to-peak signal amplitude to permit level-crossing trigger signals to be generated by an associated trigger circuit.

Thereafter, the digitized samples are analyzed to form a time histogram by counting the number of samples that occur over each of a number of signal periods for cyclic waveforms such as sine or square waves, or intervals for complex waveforms, using level-crossing signal analysis to determine the signal intervals. That is, a signal interval (which for purposes of this discussion includes signal periods) is detected when the signal passes through a first level, then a second level, and then back to the first level, such as a low-high-low sequence or a high-low-high sequence. These occurrences are ordered in the form of lengths of intervals on a time histogram to determine which interval occurs most often. The interval thus determined as being the predominant interval of the input signal may then be used to adjust the horizontal time base to result in an on-screen displayed waveform of a predetermined or user-selected number of cycles of the signal.

It is therefore one feature of the present invention to use amplitude and time histograms in analyzing an input signal in a digital oscilloscope.

It is another feature of the present invention to provide an oscilloscope with display scaling and triggering information employing histograms to extract amplitude and time information from a waveform.

Other features, and advantages of the present invention will become obvious to those having ordinary skill in the art upon a reading of the following description when taken in conjunction with the accompanying drawings.

### Brief Description of the Drawings

Fig. 1 is a generalized block diagram of a digital oscilloscope having a method of signal analysis employing histograms in accordance with the present invention;

Figs. 2A and 2B are respectively a repetitive waveform and its amplitude histogram;

Figs. 3A and 3B are respectively a cyclic waveform and its time histogram; and

Figs. 4A and 4B are respectively a complex waveform and its time histogram.

### Detailed Description of the Invention

Referring to Fig. 1 of the drawings, there is shown a generalized block diagram of a digital oscilloscope in which signal analysis in accordance with the present invention may be achieved. The circuitry is largely conventional, and therefore will be discussed in functional terms.

An input repetitive signal is applied via an input terminal 10 to a signal conditioning circuit 12. Signal conditioning circuit 12 may include conventional attenuators and gain-switching preamplifiers to adjust the input signal amplitude to a suitable level. The preamplified electrical signal is then applied to a track-and-hold/analog-to-digital converter (ADC) 14, which converts instantaneous values of the input signal to digitized samples at a rate determined by an acquisition clock 16. A replica of the input signal is applied from the signal conditioning circuit 12 to a trigger circuit 18, which issues a valid trigger signal upon the occurrence of a predetermined event (to be described later). The digitized samples produced by ADC 14 are continuously written into addresses in an acquisition memory 20 until acquisition is halted by a valid trigger event, freezing the memory contents. Thereafter, the digitized samples may be processed as digital data by a system microprocessor or CPU 22 and transferred to a system memory 24, from which such digital data may be reconstructed for display.

CPU 22, which may suitably be a conventional microprocessor and its associated kernel devices, provides overall control of the oscilloscope system, and is connected via a CPU bus 26 to acquisition memory 20, system memory 24, a user interface 30 and a display unit 32. CPU 22 operates in accordance with its own stored program instructions and user-selected control information from the user interface 30, which may suitably include a conventional front-panel control system or a keyboard, as well as a microcontroller to scan and interpret the user selections and commands to provide the user-selected control information to CPU 22. Display unit 32 may include a digital-to-analog converter (DAC) 34, a vertical amplifier 36, a time base circuit 38, a horizontal amplifier 40, and a display device 42 for reconstructing and displaying acquired waveforms. These circuits are conventional and well known to those skilled in the art, and display device 42 may be any of a number of well-known available devices having a viewing screen suitable for displaying signal waveforms and textual information, such as a cathode-ray tube or a flat-panel liquid-crystal display.

CPU 22 analyzes the waveform from the digitized samples that are being written into the acquisition memory 20 and subsequently provides information for automatically scaling the waveform by adjusting the vertical sensitivity of amplifiers within signal conditioning circuit 12, and by adjusting the time base 38 to display a predetermined or user-selected number of cycles of waveform. Information is also provided to establish stable triggering. These functions will now be described in detail.

A stream of digitized samples from ADC 14 is continuously written into acquisition memory 20, as mentioned earlier. CPU 22 analyzes the samples to develop an amplitude histogram by counting the number of occurrences of each discrete amplitude value over some finite time interval, such as one trace, or display screen width of information. That is, suppose one trace is made up of

500 digitized samples. Each sample value is used as an index in the amplitude histogram. Refer to Fig. 2A which shows a representation of an analog waveform 50, which for purposes of discussion is a sine wave. Vertical lines 52 represent the discrete amplitude values of instantaneous samples taken along the time axis of the waveform. CPU 22 determines the number of times over one trace that each discrete amplitude value occurs, and internally constructs an amplitude histogram as shown in Fig. 2B.

From the amplitude histogram, the minimum (MIN) and maximum (MAX) amplitude values of the waveform are quickly discerned, and the two values are stored. The minimum and maximum amplitude values are used to determine whether or not attenuation or gain settings need to be adjusted, and the amount of adjustment needed. CPU 22 may then apply vertical sensitivity information, such as attenuation and vertical gain, to signal conditioning circuit 12 to appropriately scale the amplitude of the input signal to a viewable level on display device 42. Alternatively, the MIN and MAX values may be displayed to a user, who can then manually adjust the vertical sensitivity controls and settings from the front panel in the conventional manner. The histogram analysis and vertical gain adjustments may be repeated until optimal signal peak-to-peak amplitude (the distance between MIN and MAX in the histogram) is attained.

The minimum and maximum values are used to calculate two levels, e.g., at the 15% and 85% points, on the waveform for later use in making time histogram calculations and for establishing level-crossing trigger events to be generated by trigger circuit 22. The amplitude histogram form of analysis is based on statistics, and therefore it is necessary that the waveform traces being analyzed not be triggered during this time. Accordingly, trigger circuit 18 is disabled during the signal analysis.

Next, CPU 22 analyzes the digital samples representing the input signal waveform to develop a time histogram by counting the number of samples that occur over each of a number of signal intervals, using level-crossing signal analysis to determine the signal intervals. With a time histogram, signal time intervals are used as the index. These time intervals are calculated in the following manner: Referring to Fig. 3A, the input signal is represented as a sine wave 50 with amplitude on the vertical axis, and time on the horizontal axis. The sine wave passes through two horizontal lines 54 and 56, which represent respectively the two aforementioned levels calculated at the 15% and 85% points between the minimum and maximum values. It should be pointed out that these percentages for the two levels 54 and 56 are not critical, and in fact may not yield satisfactory stable triggering. Therefore, other percentages, such as 25% and 75%, or 10% and 90%, may be utilized to improve performance. The time intervals are signal intervals or periods as measured beginning at both the low and high levels. The time is measured between a starting point at each level to an end point that is the same point on the

waveform as the starting point to determine the time for one interval of signal. That is, a signal interval is detected when the signal passes through a first level, then a second level, and then back to the first level, such as a low-high-low sequence or a high-low-high sequence. For example, the length of each interval of signal in Fig. 3A is represented as being equal to either the number of samples taken over the measured interval, or by the number of clock cycles that occur over the measured interval, each of which can be converted to a time interval. These interval analyses are made for every interval starting at the low levels and at the high levels, and the number of occurrences of each time interval are assembled into a time histogram as shown in Fig. 3B. For example, in Fig. 3A a length of 20 samples occurred twice and a length of 21 samples occurred once, so in Fig. 3B, which depicts the number of occurrences on the vertical axis and length on the horizontal axis, a time histogram is constructed showing two occurrences at a length of 20 and one occurrence at a length of 21. The interval that occurs most often, a length of 20 in this case, is considered to represent the interval of the signal (calculated in the above example as 0.2 microseconds). Hence, if the signal is a sine wave, the frequency of the signal may be calculated as the reciprocal of the period (calculated in the above example as 5 megahertz).

The interval determined as being the predominant interval of the input signal may then be used to manually or automatically adjust the horizontal time base to result in an on-screen displayed waveform of a predetermined or user-selected number of cycles of the signal. Suppose, for example, that the user wishes to view two cycles of a 5-megahertz signal. Since each cycle of signal has a period of 0.2 microseconds, the length of displayed waveform over ten horizontal divisions will be adjusted to 0.4 microseconds, or 40 nanoseconds per division. The time base setting may therefore be quickly calculated and selected by the microprocessor or CPU 22, or may be displayed to a user who can then manually adjust the time base controls.

CPU 22 determines either a low trigger level or a high trigger level from intervals which best match the signal period. The selected low or high trigger level is then applied to trigger circuit 12, and triggered operation is commenced. It should be noted that if neither the low level nor high level match a predominant signal interval, both a low and a high trigger level may be selected for use at the same time, providing a two-level triggering mode to obtain a stable display.

Another example of time histogram signal analysis is shown in Figs. 4A and 4B. In Fig. 4A, a more complex waveform 50 is shown along with low and high levels 54, 56 determined from an amplitude histogram analysis. Here, the intervals T0, T2 and T4 are the low-high-low transition intervals, and the intervals T1, T3 and T5 are the high-low-high transition intervals. Let us suppose that intervals T0 and T4 are the same length, and T1, T2, T3, and T5 are the same length but slightly longer than T0 and T4. These occurrences are counted and

assembled into the time histogram shown in Fig. 4B. As can be discerned from the time histogram, the interval represented by T1, T2, T3, and T5 is the predominant interval, and is used to determine the time base setting. Also, since this interval value recurs most often, level 56, in this instance, is selected as the trigger level and applied to trigger circuit 18.

Thus, it can be discerned that an amplitude histogram analysis is first used to determine minimum and maximum amplitude levels of the signal, and then a time histogram analysis is used to determine the period, or interval of the input signal. From these two analyses, vertical gain and time base adjustments or settings may be made either manually or automatically using conventional techniques, and a triggering level (or dual triggering level) established to provide a stable, usable display.

While I have shown and described the preferred embodiment of my invention, it will be obvious to those skilled in the art that many changes and modifications may be made without departing from my invention in its broader aspects. It is therefore contemplated that the appended claims will cover all such changes and modifications as fall within the true scope of the invention.

## Claims

1. A method of analyzing input signals to providing display scaling triggering for a digital oscilloscope, comprising the steps of:
  - (a) converting an input signal to digitized samples;
  - (b) assembling the amplitude values of said digitized samples into an amplitude histogram determining therefrom the minimum and maximum amplitudes of said input signal;
  - (c) establishing from said minimum and maximum amplitudes a low voltage level and a high voltage level intermediate said minimum and maximum amplitudes;
  - (d) detecting time intervals occurring in said input signal and assembling detected time intervals in a time histogram to determine a predominant time interval, wherein each detected time interval is the time for the input signal to pass through a first amplitude level, then a second amplitude level, and then back to the first amplitude level, said first amplitude level comprising one of said low voltage level and said high voltage level, and said second amplitude level comprising the other of said low voltage level and said high voltage level; and
  - (e) selecting at least one of said low voltage level and said high voltage level as a trigger level.
2. A method of analyzing an input signal to provide control information for a digital oscilloscope, comprising the steps of:

- (a) determining the maximum and minimum amplitudes of said input signal;
- (b) establishing a high threshold level and a low threshold level from said maximum and minimum amplitudes; 5
- (c) measuring a plurality of signal time intervals each representing the time taken for the input signal to pass through one of said threshold levels, reach the other threshold level and return to said one threshold level; and 10
- (d) determining from said measured intervals an interval characteristic of the input waveform and providing an output representative of said characteristic interval for use in controlling the oscilloscope time base. 15
3. A method according to claim 2, wherein said signal time interval is used to select at least one of said high threshold level and said low threshold level as a trigger level to be used to trigger the oscilloscope display. 20
4. A method according to claim 2 or 3, including the step of disabling a trigger circuit of the oscilloscope during the signal analysis. 25

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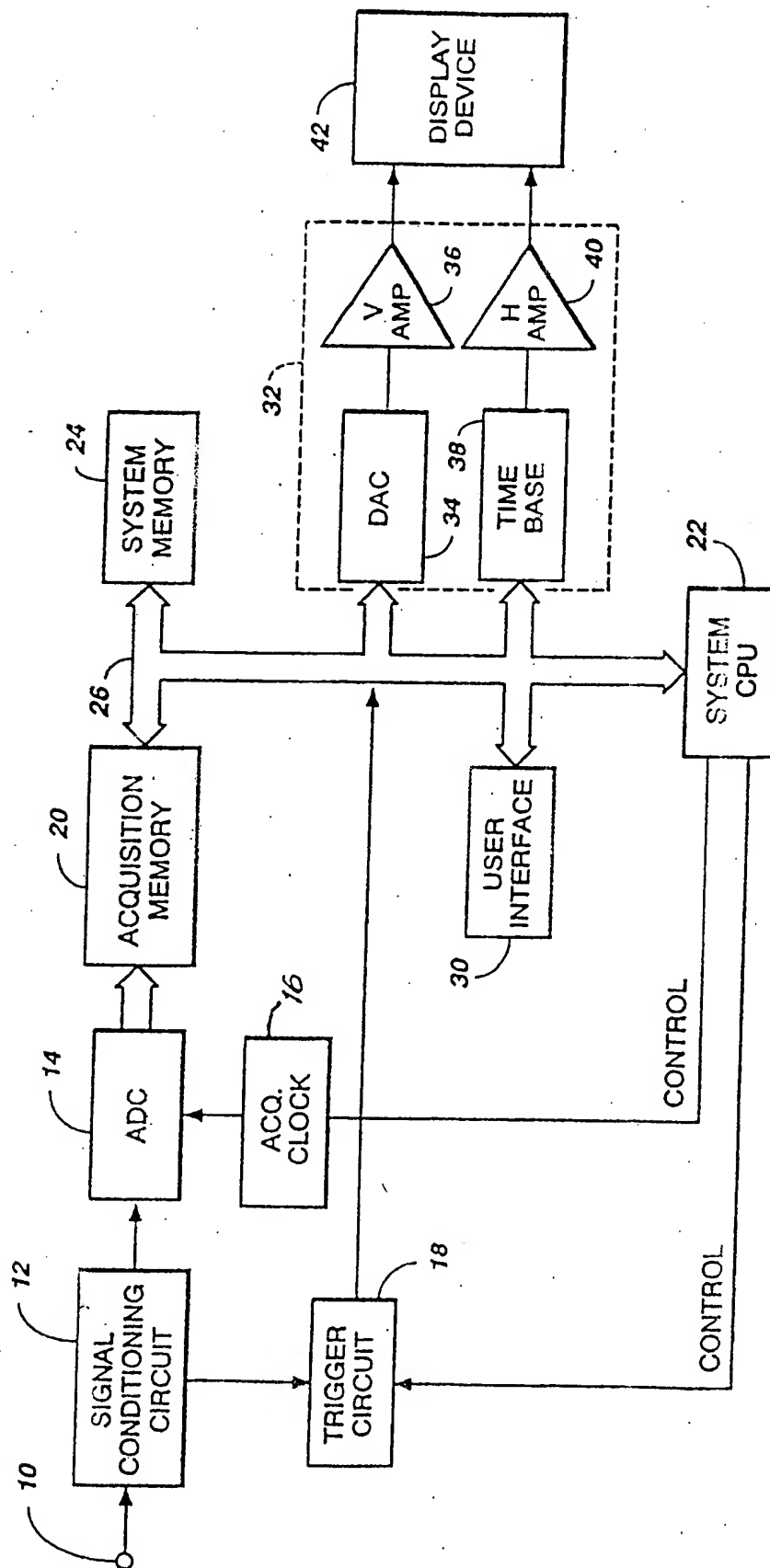
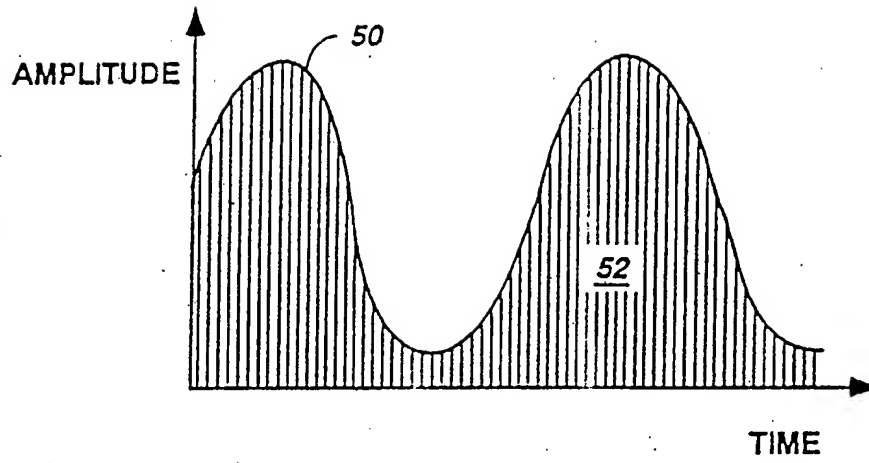
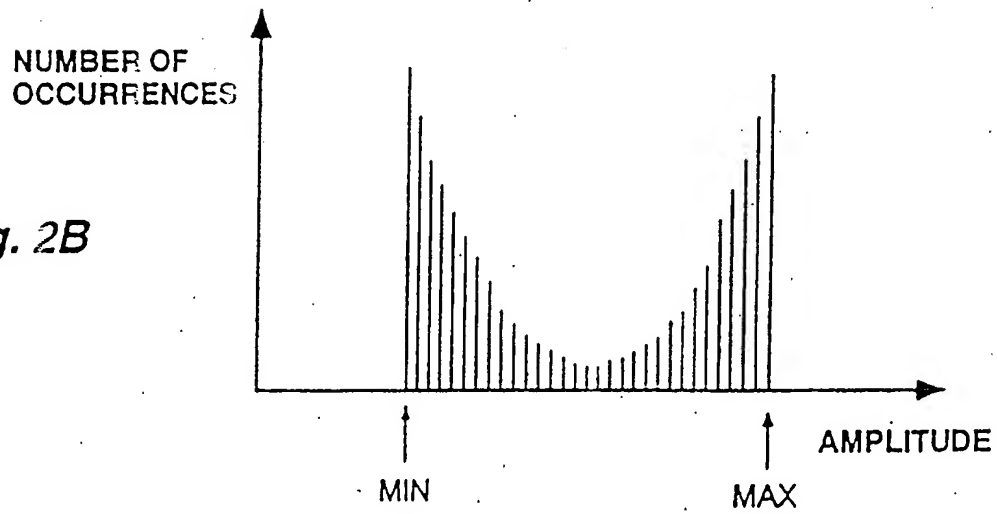


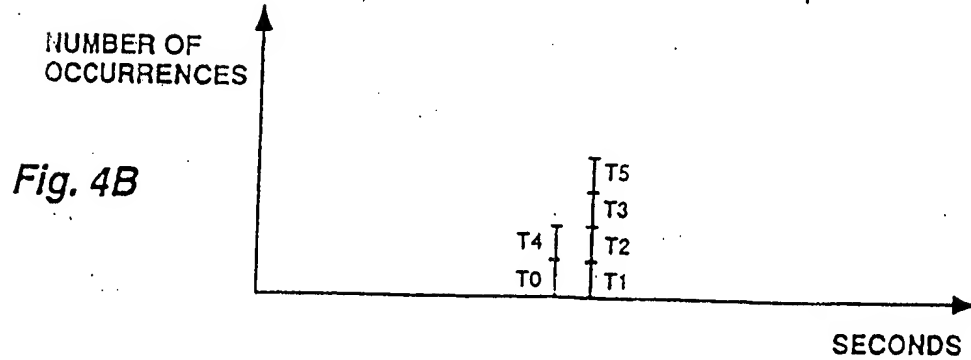
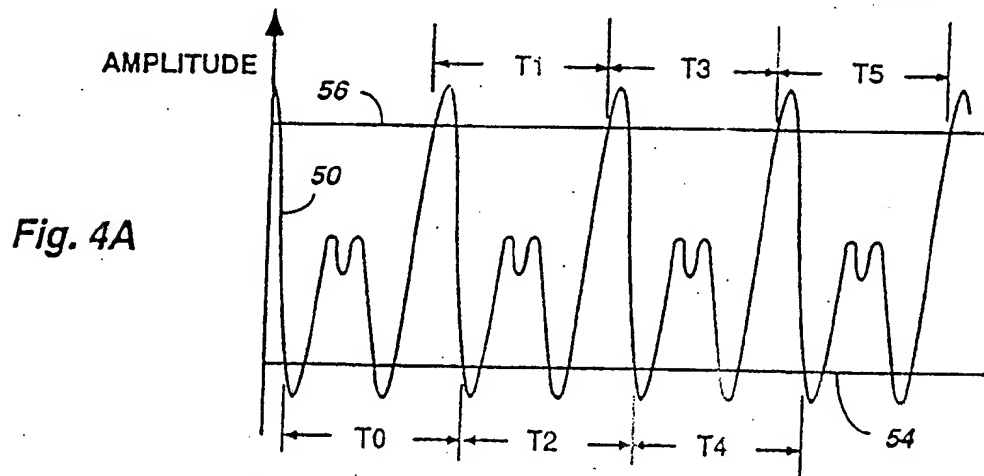
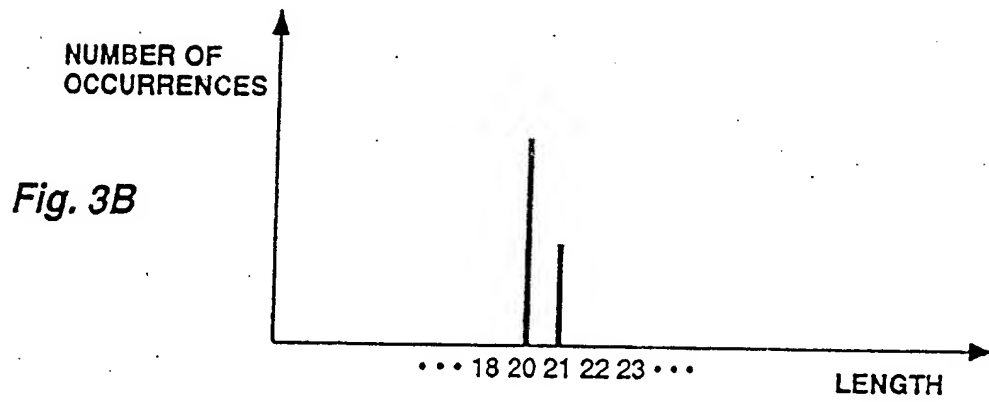
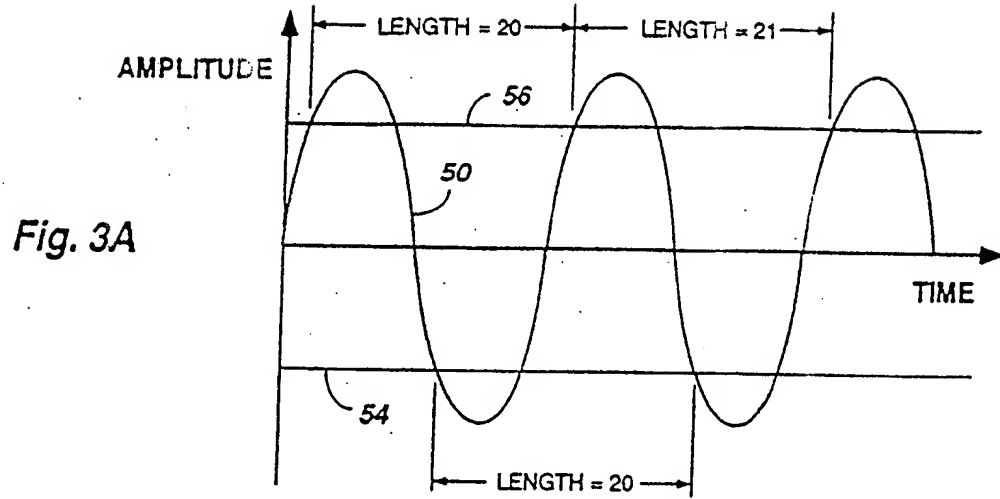
Fig. 1

**Fig. 2A**



**Fig. 2B**







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## (54) Method of signal analysis employing histograms

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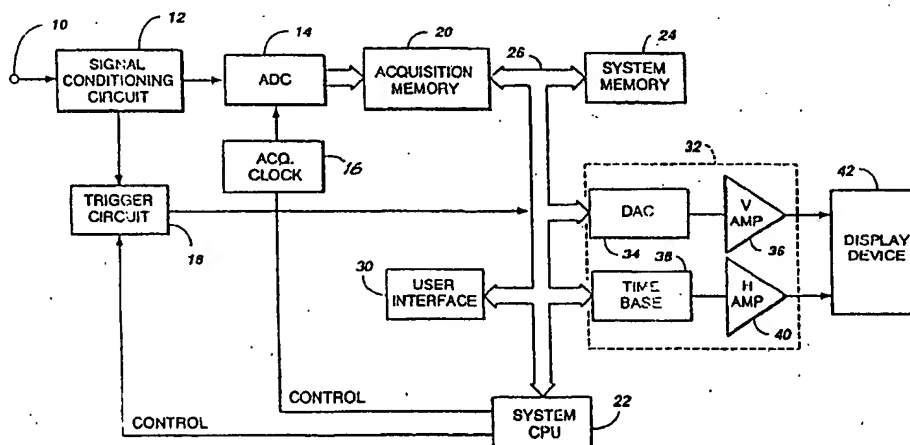


Fig. 1

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# EUROPEAN SEARCH REPORT

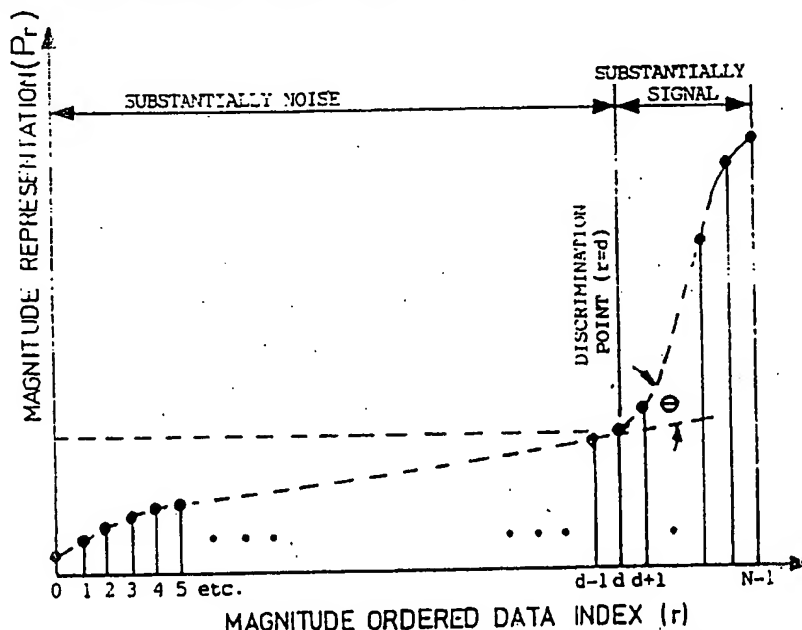
Application Number  
EP 95 30 5879

DOCUMENTS CONSIDERED TO BE RELEVANT			
Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.6)
A	EP 0 225 704 A (TEKTRONIX INC) 16 June 1987 * page 1, line 7 - line 21 * * page 6, line 34 - page 7, line 4 * * page 9, line 10 - line 19 * * figures 1,2B,3B *	1-3	G01R13/34
A	EP 0 349 328 A (TEKTRONIX INC) 3 January 1990 * abstract * * column 2, line 18 - line 24 * * column 3, line 63 - column 4, line 6 * * figure 1 *	1,2,4	
			TECHNICAL FIELDS SEARCHED (Int.Cl.6)
			G01R
The present search report has been drawn up for all claims			
Place of search THE HAGUE		Date of completion of the search 19 February 1997	Examiner Lopez-Carrasco, A
<p><b>CATEGORY OF CITED DOCUMENTS</b></p> <p>X : particularly relevant if taken alone Y : particularly relevant if combined with another document of the same category A : technological background O : non-written disclosure P : intermediate document</p> <p>T : theory or principle underlying the invention E : earlier patent document, but published on, or after the filing date D : document cited in the application L : document cited for other reasons ..... &amp; : member of the same patent family, corresponding document</p>			

## INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

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<p>(21) International Application Number: PCT/GB92/00232</p> <p>(22) International Filing Date: 7 February 1992 (07.02.92)</p> <p>(30) Priority data: 91 03931.3 26 February 1991 (26.02.91) GB</p> <p>(71) Applicant (for all designated States except US): DSP CONSULTANTS LIMITED [GB/GB]; 14 Hawthorn Road, Hatfield Peverel, Chelmsford CM3 2SE (GB).</p> <p>(72) Inventor; and (75) Inventor/Applicant (for US only): SUMMERS, Brent, Edwin [GB/GB]; 14 Hawthorn Road, Hatfield Peverel, Chelmsford CM3 2SE (GB).</p> <p>(74) Agent: LOVEN, Keith, James; Loven &amp; Co, Moulsham Mill, Parkway, Chelmsford CM2 7PX (GB).</p>		<p>(81) Designated States: AT, AT (European patent), AU, BB, BE (European patent), BF (OAPI patent), BG, BJ (OAPI patent), BR, CA, CF (OAPI patent), CG (OAPI patent), CH, CH (European patent), CI (OAPI patent), CM (OAPI patent), CS, DE, DE (European patent), DK, DK (European patent), ES, ES (European patent), FI, FR (European patent), GA (OAPI patent), GB, GB (European patent), GN (OAPI patent), GR (European patent), HU, IT (European patent), JP, KP, KR, LK, LU, LU (European patent), MC (European patent), MG, ML (OAPI patent), MN, MR (OAPI patent), MW, NL, NL (European patent), NO, PL, RO, RU, SD, SE, SE (European patent), SN (OAPI patent), TD (OAPI patent), TG (OAPI patent), US.</p> <p>Published With international search report.</p>

(54) Title: SIGNAL PROCESSING APPARATUS AND METHOD



(57) Abstract

The invention provides a real-time (or non real-time) signal processing apparatus which enables the accurate and robust discrimination, separation and suppression (or detection) of one or more component random signals/variables which comprise an unknown proportion of a composite random signal (or statistical mixture/ensemble) contained within a finite bandwidth; wherein, the component random signals/variables are characterised by having different (and not necessarily known) statistical attributes, e.g. cumulative probability distribution functions. The signal processing method and apparatus, which are suitable, for example, for audio noise reduction, divide a digital signal into successive sets of data and derive from each set a power spectrum. By sorting the power spectrum into rank order of magnitude, a point ( $P_d$ ) along the rank ordered data set ( $P_r$ ) where a change of gradient or rate of change of magnitude with element number is detectable in the case where the signal is a composite signal containing both signal and noise, and the elements having the lower gradient represent the noise component. The original signal can therefore be processed to suppress or zero those elements which have been identified as noise.

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SIGNAL PROCESSING APPARATUS AND METHOD

This invention relates to a signal processing apparatus and method for use in discriminating between the component random signals, having different statistical distributions, in a composite signal consisting of a summation of two or more such random signals.

The invention enables the accurate and robust discrimination, and preferably separation and suppression (or detection), of one or more component random signals. One preferred embodiment of the apparatus and method is for example for improving the signal-to-noise ratio of a band limited signal, such as audio.

For an audio noise reduction system, advantage may be taken of the inherent spectral redundancy associated with audio signals. The audio frequency spectrum is divided into spectrally adjacent frequency bands which are independently deactivated (or suppressed) according to the instantaneous spectral occupancy of the audio signal. By deactivating the frequency bands which do not contain signal, net bandwidth reduction occurs resulting in an overall improvement in signal-to-noise ratio (S/N). The "quality" of such a system is essentially determined by the accuracy with which the noise can be discriminated from the signal. Therefore, if the frequency bands containing signal are mistaken for frequency bands containing noise, their subsequent rejection or suppression will result in audible distortion. Whereas,

in the event that frequency bands containing noise are mistaken for frequency bands containing signal, their inclusion will degrade the potentially available S/N improvement.

In general, both audio signals and noise may be considered as random signals. Therefore, the composite signal resulting from audio plus noise, i.e. noisy audio, may also be considered as a random signal consisting of the sum of two random signals from different and generally unknown (especially in the short term) statistical distributions. Thus, the primary mission of an audio noise reduction apparatus is to accurately separate the two random signal components from an unknown statistical mixture, and then remove or suppress the noise component to give a noise-reduced signal.

It has now been found that if the results of a N-point discrete transform of such a sampled composite signal or statistical mixture (e.g. the power spectra of a temporal section of a noisy audio signal) are considered as a set of random numbers and are subsequently arranged in magnitude (or rank) order, as illustrated in Figure 1 for example, there is a certain point (d) along the rank ordered data set where a noticeable change in the rate of change of magnitude difference between adjacent data elements (or gradient) is evident. For a signal composed of the summation of two random signals from different statistical dis-

tributions, it has been realised that this point in the data set represents the discrimination point between elements of a component signal substantially from one distribution and those from the other. For an ascending order ranking, data elements to the left of the point are largely from one component signal (e.g. noise), and data elements to the right of the point, which exhibit a higher gradient, are substantially from the other component signal, e.g. audio.

The invention provides a real-time (or non real-time) signal processing apparatus which enables the accurate and robust discrimination, separation and suppression (or detection) of one or more component random signals/variables which comprise an unknown proportion of a composite random signal (or statistical mixture/ensemble) contained within a finite bandwidth, wherein the component random signals/variables are characterised by having different and not necessarily known statistical attributes, e.g. cumulative probability distribution functions. Preferably, the invention achieves the aforementioned by providing means for accepting an already digitised composite signal, or alternatively for digitising an incoming composite analogue signal, such that the samples are a 'unique representation' (in the sampling theorem sense, i.e. taken at a rate appropriate to the signals bandwidth), and also represented by an appropriate level of quanta, means for collecting (overlapped) batches of N successive signal samples, means

for storing the batch of N signal samples for future use, means for transforming each batch of the said signal samples into a 'discrimination space' (with appropriately controlled sidelobes) such as for example, the 'frequency domain' by means of a weighted Discrete or Fast Fourier Transform or some other transform such as a Cosine, Hartley, Bracewell or Walsh, means for computing a representation of the magnitude of the signal representation within the discrimination space, e.g. the modulus (or power spectrum), means for sorting the discrete values of the signal magnitude representation within the discrimination space (e.g. the line spectra) into ascending or descending rank order, means for identifying a discrimination point (or points) within the said rank ordered data set, preferably by identifying a point or points at which there is a significant change in gradient, and means for separating the data elements to one side (or in between) of the identified discrimination point (or points) whereby the data elements to one side (or in between) of the discrimination point (or points) substantially represent elements comprising one of the component signals.

Means may also be provided for identifying and retaining or for identifying and suppressing those data elements to one side of, or in between, the discrimination point (or points) in the original un-sorted data set representing the elements of a component signal (or signals)



within the composite signal represented in the discrimination space. Further means may output the data elements representing a component signal (in the case of a signal detection type system) or use the data as a basis for the real-time design of an adaptive filter which is subsequently applied to the original (but delayed) batch of N signal samples previously stored yielding a component signal suppressed output (eg noise suppressed) means for reconstructing the processed output samples back into analogue form or means for outputting the processed data directly in digital form. The aforementioned method performed by the apparatus is repeated on successive blocks of data with preferably 50% or more block overlap, and means may be provided for removing data block end effects (by, for example, interpolation or incorporating an attack and decay function on the filter amplitudes block to block).

Preferably, the method of the invention comprises accepting an already digitised composite signal, or alternatively sampling an incoming composite analogue signal such that the samples are a 'unique representation' (in the sampling theorem sense, i.e. taken at a rate appropriate to the signals bandwidth), and also represented by an appropriate level of quanta, and then collecting (overlapped) batches of N successive signal samples, storing the batch of N signal samples for future use, performing a transform operation on each batch of the said signal samples such that

the composite signal is represented in some new 'discrimination space' (with appropriately controlled sidelobes) such as for example, the 'frequency domain' by means of a weighted Discrete or Fast Fourier Transform or some other transform such as a Cosine, Hartley, Bracewell or Walsh, calculating a representation of the magnitude of the signal representation within the discrimination space, e.g. the modulus (or power spectrum), sorting the discrete values of the signal magnitude representation within the discrimination space (e.g. the line spectra) into an ascending or descending rank order, identifying a discrimination point (or points) within the said rank ordered data set, preferably by identifying a point or points at which there is a significant change in gradient, separating the data elements to one side (or in between) of the identified discrimination point (or points) whereby the data elements to one side (or in between) of the discrimination point (or points) substantially represents the data elements comprising one of the component signals, identifying and retaining or identifying and suppressing those data elements to one side of the discrimination point (or points) in the original unsorted data set representing the elements of a component signal (or signals) within the composite signal represented in the discrimination space, outputting the data elements representing a component signal (in the case of a signal detection type system) or using the data as a basis for the

design of an adaptive filter which is subsequently applied to the original (but delayed) batch of N signal samples previously stored yielding a component signal suppressed output e.g. noise suppressed, and reconstructing the processed output samples back into analogue form or outputting the processed data directly in digital form. The aforementioned method is repeated on successive blocks of data with preferably 50% (or more) block overlapping and may suppress data block end effects, for example by interpolation or incorporating an attack and decay function on the filter amplitudes block to block.

Preferably, the step of determining the discrimination point from the rank ordered sequence of the discrimination space representation is achieved by means of calculating an amplitude independent 'discrimination sequence or function' such as, for example,

$$g(d) = \frac{d \cdot P_{(d)}}{\sum_{r=0}^d P_{(r)}} \quad ; d = 0, 1, \dots, N-1 \quad (1)$$

where  $P(r)$  is the rank ordered data sequence.

Having computed (or in the process of computing)  $g(d)$ , a value of  $d$  can be determined such that  $g(d)$  is equal to (within some specified bounds) a predetermined value and  $g(d+1)$  is greater than said predetermined value. For many circumstances, it has been found that, in the case of two component signals, a value of  $g(d)$  in the range of 1.9 to

2.4 yields an accurate estimate of the discrimination point (where the sorted data sequence are magnitude values and  $d$  is at least 20 to overcome any initial small data set problems). However, depending on the application, other values of  $g(d)$  may also be selected which may be more appropriate to the nature of the signals being analysed and/or the specific application to which the method is being applied.

For the majority of applications, the accuracy with which the method described hereinbefore determines the discrimination point will be sufficient. However, in order to obtain even greater separation accuracy it has also been realised that the angle  $\theta$  (or some equivalent measure thereof) subtended between the two gradients either side of the discrimination point - see Figure 1 - is a direct measure of the 'difference' between the two statistical distribution functions from which the two component signals can be considered as originating. In other words, it is a direct measure of "statistical confidence" in the particular data set being analysed.

Since ultimately a decision has to be made as to which data elements are to be declared as belonging to a particular component signal (or distribution), the statistical confidence factor can be used to bias or weight the decision made. Thus, if  $\theta$  tended towards  $90^\circ$ , this would signify that the two data sets originate from very different

statistical distribution functions and consequently the confidence in the decisions made about which elements belong to which component signal (or distribution) would be high, i.e. tending towards 100%. If however,  $\theta$  tended towards  $0^\circ$ , then this would signify that the two data sets originate from very similar statistical distribution functions and confidence in the subsequent decisions would be very low, i.e. tending towards 0%. Since in audio noise reduction systems for example it would be generally preferable to degrade the potentially available S/N improvement rather than cause audible distortion, the lower the confidence in a particular set, the more it might be desirable to bias the discrimination point towards the left, i.e. to discard fewer of the supposed noise samples. If no discrimination point has been detected, i.e.  $\theta = 0^\circ$ , then this implies that only one distribution is present, or that if two distributions are present, it is impossible to discriminate between them.

To explain this further, suppose a temporal section of a composite signal comprising signal plus noise and having a positive signal-to-noise ratio is sampled, weighted, Discrete Fourier transformed and modulus-squared in order to obtain the power spectrum. In addition, suppose that the amplitude of the composite signal is adjusted such that, regardless of the temporal section selected, the peak value of the composite power spectrum is never more than the value  $M$ . If the probability density function of the signal plus

noise were plotted, then it would have, for example, a 'normal distribution' centred on  $M/x$ , as shown in Figure 2(a). That is to say, the magnitudes of the signal power spectra would be most likely to have a value near  $M/x$ , whereas the likelihood of their having a value of 0 or  $M$  is small by comparison. Since the composite signal has a positive signal-to-noise ratio, the mean noise magnitude will be located at some lower value, say  $M/n.x$ , where  $n$  is some positive constant. If the probability density function for the noise alone were plotted, it would have, for example, a Rayleigh distribution with mean  $M/n.x$  as also shown in Figure 2(a). It will be seen that magnitude  $P_{(r-d)}$  is the magnitude of the intersect point of the two distributions, and defines that magnitude of the composite power spectrum which is equally likely to belong to either distribution, and corresponds directly to the gradient transition magnitude in the rank ordered sequence hereinbefore described. Similarly, Figure 2(b) illustrates the cumulative distribution functions (cdf) for the same scenario. It can be seen from this diagram that the gradient transition point in the rank ordered data sequence is also equivalent to the magnitude at which there is maximum difference between the cdf of noise alone and cdf of signal plus noise. This diagram in particular is indicative of why the method herein described is so robust. If magnitude  $P_{(r-d)}$  is declared as the threshold value, then everything to the left of the line

in the case of an ascending ranking is deemed to be noise and as such is either discarded or suppressed, leaving what is substantially signal to the right of the line. However, it can be seen from Figure 2 that in both sections of the density functions either side of the threshold value there is a residual component of the other distribution. This is an inevitable consequence of any thresholding system, as is well known.

For the case where the system is a noise rejection system, it is possible to define two statistical noise rejection parameters, i.e. the probability of detecting noise ( $P_{dn}$ ), and the probability of falsely declaring signal as noise ( $p_{fs}$ ).  $P_{dn}$  is specified as a percentage, e.g. 90%, whereas  $p_{fs}$  can be specified either as a percentage or as a ratio. Here we shall adopt the ratio notation e.g.  $10^{-6}$ , read as "one noise declaration in a million will be a falsely declared signal". With reference to Figure 2(a),  $P_{dn}$  and  $p_{fs}$  are equivalent to the areas under the curves as indicated by the shadings (Note that the total area under each curve is normalised to unity). In the same way that signal detection systems trade between the probability of detection and the probability of a false alarm ( $p_{fa}$ ), final selection of a threshold is a trade-off between  $P_{dn}$  and  $p_{fs}$ . For example, it can be seen from Figure 2(a) that if the threshold line at magnitude  $P_{(r-d)}$  were moved towards the right of the illustration,  $p_{fs}$  will increase more rapidly than  $P_{dn}$ . That

is to say, in order to increase the probability of correctly detecting noise, the probability of falsely declaring a signal as noise also increases, but at a much faster rate. It can be seen therefore that for optimum performance the correct and robust choice of the threshold point is critical.

As illustrated, Figure 2 implies that the probability density functions of both signal plus noise and noise alone are well-behaved. Such circumstances are only true of long-term statistics, i.e. where the observation time of the sample set is very large. For practical systems, it is necessary to process relatively small data sets. However, the problem with doing this is that short term statistics are notoriously unreliable, in that the probability density functions distort in an unpredictable manner, due to insufficient data, making it sometimes impossible to discriminate between them. Consequently, this results in a situation wherein the accuracy of, and the confidence in, any decision made about the data set is very low.

The rank ordering discrimination process as hereinbefore described includes within the method of the invention a means of extracting an inherent and direct measure of confidence in the data set being analysed. This confidence factor can therefore be used, in the case of audio noise reduction for example, to bias the decision point adaptively in favour of reduced S/N improvement for situations of low confidence, such as poor S/N.



The number of samples processed in a particular batch (N) is preferably not less than 128 for reliable discrimination. It should also be noted that, since the number of points in a Transform process (as well as the window function selected) determines the process coherent gain, the system will work with input signals having negative signal to noise ratio, i.e. with more noise than signal.

The apparatus and method of the invention enable the accurate and robust discrimination, separation and suppression (or detection) of one or more component random signals/variables which comprise an unknown proportion of a composite random signal (or statistical mixture/ensemble) contained within a finite bandwidth; wherein, the component random signals/variables are characterised by having different (and not necessarily known) statistical attributes, e.g. probability density functions. It has been found that the apparatus is particularly suitable for real-time implementation as the method exhibits robust performance hitherto unattainable by more well-known methods in terms of its separation accuracy under a wide range of input signal conditions. In addition, it has also been found that the method maintains its accuracy performance even with small (in statistical terms) sample sets. It is envisaged that this invention will find application in most types of signal rejection and signal detection systems particularly when accuracy of the detection/rejection is of prime importance.

It will be appreciated that, since the apparatus and method of the invention process discrete temporal sets of digital samples selected from some arbitrary starting point, repeating the process with a different starting point would afford a further reduction in noise. Thus, a two or more stage system could also be envisaged, although the additional S/N improvement of each additional stage would be increasingly small, since each set would become increasingly correlated. Similarly, multi-channel versions can be envisaged by simple repeats of the process herein described.

Reference is now made to the drawings, in which:

Figure 1 illustrates the rank ordered power spectra of a sampled temporal section of a composite signal comprising two component random signals from different probability density functions such as audio and noise;

Figure 2(a) illustrates the relationship between the discrimination point and typical probability density functions of signal plus noise and noise alone such that there is a positive S/N;

Figure 2(b) illustrates the relationship between the discrimination point and typical cumulative distribution functions and signal plus noise and noise alone such that there is positive S/N;

Figure 3 is a block diagrammatic representation of an apparatus according to one exemplary embodiment of the invention for use as a real-time audio noise reduction system, i.e. an adaptive spectral noise gate;

Figure 4 is a graph corresponding to that of Figure 1, but illustrating the effect of reducing, rather than completely suppressing, the noise content of the signal;

Figure 5a illustrates the instantaneous power spectrum of some hypothetical temporal section of an audio plus noise signal;

Figure 5b illustrates the frequency response of a filter designed for such a signal;

Figure 5c illustrates the result of multiplying the original signal power spectrum with the frequency response of the filter;

Figure 6 is a flow chart illustrating the method by which the threshold level may be determined;

Figure 7 is a graph illustrating the ideal nth filter amplitude profile; and

Figure 8 is a flow chart illustrating a method of adjusting the ideal filter levels.

Reference is now made to Figure 3, which illustrates the data-flow diagram of an apparatus according to one exemplary embodiment of the invention for use as a real-time

audio noise reduction system. The apparatus illustrated in Figure 3 will now be explained in full detail with reference also to Figures 4 to 8.

N noisy audio signal samples are collected in a double buffered store 8 via an input selection switch 7 from either an analogue input 1 via a buffer amplifier 3, lowpass filter 4 and analogue to digital converter 5, or a digital audio input 2 comprising a digital audio interface 6 (e.g. AES/EBU or S/PDIF). During the time  $\tau$  it takes to collect the next batch of noisy audio samples, where

$$\tau = N \cdot \alpha \cdot T \quad (2)$$

and  $N$  = Number of samples (typically 128)

$\alpha$  = Block overlap ratio e.g. 0.5 for 50%

$T$  = time interval between samples (must satisfy Nyquist) typically  $1/(48 \text{ kHz})$  or  $1/(44.1 \text{ kHz})$  for audio,

the current batch 'b' of data  $\{\{x_{kT+bT}\}_{k=0}^{N-1}\}_{b=0}^{\infty}$  (where 'k' is

the sample index and 'b' is the batch index) is passed to a memory device 11 which stores the data for future use and a processing device 9 which weights the data with a window function 10 to produce the output sequence  $\{\{z_{kT+bT}\}_{k=0}^{N-1}\}_{b=0}^{\infty}$  where

$$\{\{z_{kT+bT}\}_{k=0}^{N-1}\}_{b=0}^{\infty} = \{\{x_{kT+bT} \cdot w(l)_{kT}\}_{k=0}^{N-1}\}_{b=0}^{\infty} \quad (3)$$

and  $\{w(l)_{kT}\}_{k=0}^{N-1}$  = Pre-computed N-point window function such as a -90dB Dolph-Chebyshev or similar. Note -90dB

sidelobe-level is an arbitrary choice and although typical can be selected according to the signal quality processed.

The weighted data sequence is then transformed at 12 into the discrimination space by means of, for example, a Fast Fourier Transform (or similar) to produce (in the case of the FFT/DFT) the complex-valued N-point data sequence  $\left\{ \left\{ a_{\frac{n}{NT}, b} + jA_{\frac{n}{NT}, b} \right\}_{n=0}^{N-1} \right\}_{b=0}^M$  where (in the case of the FFT/DFT) for the b-th batch

$$\left\{ a_{\frac{n}{NT}, b} + jA_{\frac{n}{NT}, b} \right\}_{n=0}^{N-1} = \left\{ \sum_{k=0}^{N-1} z_{kT+bT} \cdot e^{-j \frac{(2\pi nk)}{N}} \right\}_{n=0}^{N-1} \quad (4)$$

The power spectra  $\left\{ P_{\frac{n}{NT}, b} \right\}_{n=0}^{N-1}$  for the b-th batch is then obtained by computing the magnitude-squared at 13. That is

$$\left\{ P_{\frac{n}{NT}, b} \right\}_{n=0}^{N-1} = \left\{ a_{\frac{n}{NT}, b}^2 + A_{\frac{n}{NT}, b}^2 \right\}_{n=0}^{N-1} \quad (5)$$

Alternatively, the magnitude sequence could be computed, but this is generally a more computationally intensive operation to perform for no additional benefit in this case.

The power spectra of the b-th batch are then passed to a further computation stage 14 which sorts the sequence by amplitude into ascending rank-order. (Note that since the original data sequence was real-valued it is only necessary to operate on the first-half (plus one) of the data array since the second-half is merely the image of the first and as such will reveal no further information. This is a highly significant observation for performing data sorts in real-time. However, if advantage of this fact is

taken then it is important to note that the value  $P_{n-0}$  is replaced with the value  $P_{n-0}/4$  and similarly  $P_{n-\frac{N}{2}}$  is replaced with the value  $P_{n-\frac{N}{2}}/4$  just before sorting for proper scaling.) Then, for the b-th batch

$$\left\{ P_{\frac{r}{NT}, b} \right\}_{r=0}^{\frac{N}{2}} = \text{SORT} \uparrow \left[ \left\{ P_{\frac{n}{NT}, b} \right\}_{n=0}^{\frac{N}{2}} \right] \quad (6)$$

Note that since the values of the sequence elements have not altered through this process, but only their indices, the original index 'n' has been replaced by the index 'r' once sorted.

The sorted or rank ordered sequence is then passed to a further computational element 15 which finds an appropriate threshold. In order to explain this it is necessary to digress temporarily to consider the impact of equation (1) and the discrimination process herein described in this audio noise reduction application.

From equation (1) and Figure (1) it can be seen that at the discrimination point 'd'

$$\frac{d \cdot P_d}{\sum_{r=0}^d P_r} \leq \text{Constant} > \frac{(d+1) \cdot P_{(d+1)}}{\sum_{r=0}^{d+1} P_r} \quad (7)$$

where Constant = 4 (typically, for magnitude-squared sorted data)

If data elements in the sorted array to the left of the discrimination point (for an ascending order ranking)

are those substantially from noise alone, then the mean noise level may be estimated as

$$mnl = \frac{\sum_{r=0}^d P_r}{d+1} \quad (8)$$

Consequently, the instantaneous signal-to-noise ratio (SNR) of the temporal section of audio may be estimated as

$$SNR = 10 \log \left( \frac{\sum_{r=d+1}^{\frac{N}{2}} P_r}{mnl \cdot \left(\frac{N}{2}\right)^2} \right) \text{ dB} \quad (9)$$

Note that in Equation (9) the average of the sum of the noise components is assumed to be present in the frequency bands containing signal, i.e. those to the right of the discrimination point.

From the point of view of audio it has been found that to discard all of the potentially available noise components every block is too severe since the noise content can change quite dramatically in accordance with the instantaneous spectral occupancy of the signal - resulting in what is commonly termed a 'noise pumping effect'.

Consequently, it has been found that a more pleasing audio effect is obtained if the user specifies a total noise power value ' $P_r$ ' which they wish to discard or suppress. So long as this value is less than that implied by the discrimination point the noise components can be safely discarded with the assurance that no signal is being rejected.

In this way the noise-pumping effect can be avoided.

It has also been found that rather than suppress the declared noise components completely, for certain types of audio signal where the generally perceived view is that the noise is part of the 'expected' audio content, e.g. a live recording where a portion of noise adds to the atmosphere of the recording as compared to a studio recording, there are practical advantages in allowing the user to control the level of suppression.

Figure 4 illustrates the implied consequence on the sorted data sequence of performing this action. Also from Figure 4, it can be seen that

$$SNR \text{ after processing} = 10 \cdot \log \left( \frac{\frac{2}{N} \cdot \sum_{r=d+1}^{\frac{N}{2}} P_r}{MA \cdot \sum_{r=0}^u P_r + \sum_{r=u+1}^d P_r + mnI \cdot \left(\frac{N}{2} - d - 1\right)} \right) \text{ dB} \quad (10)$$

where

$$0 \leq u \leq d < \frac{N}{2}$$

MA = maximum attenuation level (user specified e.g. 0.01)

u = user implied specified point from  $P_T$ , where  $u \leq d$

Note that the user specifies  $P_T$  where

$$\sum_{r=0}^u P_r \leq P_T < \sum_{r=0}^{u+1} P_r \quad (11)$$

from which the index 'u' can be found and checked to ensure that it is less than or equal to 'd'. If not, then only the data elements up to 'd' are permitted to be suppressed by a maximum amount of MA.



From equations (9) and (10) it can be seen therefore that the signal-to-noise improvement of the process can be expressed as

$$\text{SNR improvement} = \text{SNR after processing} - \text{SNR before processing} \quad (12)$$

It should be noted, however, that whilst in the scientific sense a SNR improvement can be calculated by an equation such as (12), in terms of its audio performance, a perceived further improvement to the listener is apparent due to what is commonly termed the "masking effect".

Figure 5(a) illustrates the instantaneous power spectrum of some hypothetical temporal section of audio plus noise. Figure 5(b) illustrates the frequency response of the filter we would wish to design for such a signal. Figure 5(c) illustrates the result of multiplying the original signal power spectrum with the frequency response of the filter.

It can be seen from Figure 5(c) that the inter-line frequency bands containing noise have been suppressed whereas the noise in close proximity to a signal component has not. Therefore, whilst there is an overall signal-to-noise improvement as described by equation (12) it is not in the mean-level sense. This fact is important since it has been shown by a number of workers in the audio field that a low-level signal (such as noise in this case) cannot be

detected by the ear if it is in the close proximity to a stronger tone. In other words the close-in noise is 'masked' by the presence of the audio signal component. Because of this, the "apparent" SNR improvement of such a process may be more accurately described by

$$\text{Apparent SNR improvement} = 10 \cdot \log \left( \frac{\frac{2}{N} \cdot \sum_{r=d+1}^{\frac{N}{2}} P_r}{MA \cdot \sum_{r=0}^u P_r + \sum_{r=u+1}^d P_r} - \frac{\frac{2}{N} \cdot \sum_{r=d+1}^{\frac{N}{2}} P_r}{\sum_{r=0}^d P_r} \right) \text{ dB} \quad (13)$$

Note that unlike equation (12) if  $u = d$  and  $MA = 0$  then from equation (13) the 'effective' SNR improvement =  $\infty$ , i.e. no residual noise is audible at all!

Figure 6 illustrates one such example of the 'Find Threshold' process 16 in detailed flow-graph form.

It will be seen from Figure 6 that the find threshold process essentially consists of two parts. The first part finds the index value 'u' which corresponds to the user's input parameter ' $P_T$ ' for total noise power to be removed in accordance with equation (11). (Note that  $A = \sum_{x=0}^r P_x$ , i.e. the sum of noise components up to and including the r-th.) The second part checks the determined value 'u' against the discrimination point 'd' and backs off the index if  $u > d$  or leaves it unmodified if  $u \leq d$ . The threshold value is then equal to the amplitude pointed to by the index 'u' (or 'd' if  $u > d$ ). In other words, the threshold value is equal to either  $P_{r-u}$  or  $P_{r-d}$  if  $u > d$ . In this exemplary embodiment, the pre-determined value = 4 (see Equation (1)).

It will be appreciated that the method described for determining a threshold value on the basis of the sorted data set and discrimination point is one exemplary embodiment of the invention which yields exceptionally good results in this audio case. Clearly, depending on the application there may be many more ways of using the invention, either directly or indirectly.

Having determined a threshold value it can be seen from Figure 3 that the unsorted data sequence of the b-th batch  $\{P_{\frac{n}{NT},b}\}_{n=0}^{N-1}$  is compared with the threshold value in comparison stage 17. If the nth data element is greater than the threshold value the output of the comparison stage 17 produces a '1'. Similarly if the nth data element is less than or equal to the threshold value the output of the comparison stage 17 produces a '0' so that the 1's and 0's vector  $\{T_{\frac{n}{NT},b}\}_{n=0}^{N-1}$  is produced. That is, for the b-th batch,

$$\text{IF } P_{\frac{n}{NT},b} > \text{Threshold value THEN } T_{\frac{n}{NT},b} = 1 \text{ ELSE } T_{\frac{n}{NT},b} = 0 \}_{n=0}^{N-1} \quad (14)$$

Clearly, a value of '1' signifies that a signal component is present in that particular frequency band, whereas a '0' signifies that there is only noise. Therefore, the 1's and 0's vector can be considered as the 'ideal' sampled frequency response of the filter to be designed to pass signal but to suppress noise such that a '1' = a passband and '0' = a stopband.

Given an ideal filter sampled frequency response it is a relatively straightforward matter to obtain the filter coefficients. However, in the case of audio as so far described there is a further complication resulting from batch-to-batch effects. This audible effect is termed the "gurgling effect" as the resultant sound resembles that of a trickling stream of water. It stems from the fact that batch-to-batch both the quantity and spectral disposition of the suppressed frequency bands are completely random. It is hypothesised that if the process herein described were to be performed on every incoming sample, i.e. a running block of  $N$  rather than overlapped block processing, the gurgling effect would sound like noise again since the rate of change of the residuals is higher than the highest frequency in the audio band. However, given that the algorithm can be performed on blocks overlapped by as little as 50% the rate of change of the residuals is within the audio band and thus can be heard as an effect other than noise, i.e. a self generated signal.

It has been found that in order to suppress this residual sound, a very effective solution is to introduce a filter 'attack' and 'decay' function block-to-block. The mechanism of the 'attack' and 'decay' function will be described in detail shortly. However, before doing so, it

is appropriate at this stage to mention a further audible effect which can also be mitigated within the filter design process.

Suppose as a result of the previous analysis a noise spectral spike or short-term spectral transient is declared as 'a signal' which requires passing. The resultant audible effect termed "popping effect" sounds like a wave crashing on the rocks in that there is a short-term burst of narrow-band noise. As it is not in close proximity to a true signal component it is therefore not masked and is consequently quite audible. Although the probability of occurrence by virtue of the detection process is very low, it does nevertheless occur from time to time, since the probability of falsely declaring noise as signal ( $p_{fa}=1-P_{dn}$ ) is never zero.

In order to overcome this, a further 'temporal duration check' can be included at this point. That is to say, the probability of some audio signal component occurring for one block time only (1.3 ms if  $N=128$ ,  $f_s=48$  kHz and 50% block overlap) is very low. In this sense, even though the process is being applied in blocks rather than every sample, the process is over-sampled. Consequently, a very effective mitigation technique is to simply check that the signal exists for a number ( $C$ ) of consecutive blocks.

Typically,  $C=4$  for  $N=128$ ,  $f_s=48$  kHz and 50% block overlap, implying that a further acceptance test of the declared signal being truly signal is that it exists in the same spectral location for at least 5 ms or so. Whilst this may not sound like a very long time in audio terms, from a detection point of view this simple method dramatically decreases the probability of falsely declaring noise as signal because the pfa then becomes the cumulative probability of 'C' false declarations in succession. If, for example, the pfa was  $10^{-5}$ , i.e. one falsely declared noise component as signal every 100,000 decisions, then at 48,000 decisions per second it would be expected that the popping effect would be audible approximately every 2 seconds, which would clearly be unacceptable.

Given that there are  $\frac{N}{2}$  unique frequency bands, then

$$\begin{array}{l} \text{probability of falsely declaring} \\ \text{noise as signal in one batch} \end{array} = \frac{N \times pfa}{2} \quad (15)$$

$$\begin{array}{l} \text{probability of falsely declaring} \\ \text{noise as signal in two successive} \\ \text{batches in the same frequency band} \end{array} = \frac{N \times pfa^2}{2} \quad (16)$$

$$\begin{array}{l} \text{probability of falsely declaring} \\ \text{noise as a signal in 'C' successive} \\ \text{batches in the same frequency band} \end{array} = \frac{N \times pfa^C}{2} \quad (17)$$

In the previous example if  $pfa=10^{-5}$  and  $C=4$  then from equation (17) it can be seen that it would be expected to occur once every  $10^{20}$  decisions. At 48,000 per second, this is equivalent to  $6.6 \times 10^7$  years. Clearly, the pfa has been



to the current and previous 'C' blocks. Similarly, the amplitude of the filter decays (if subsequent signal detections reveals no signal) over subsequent blocks by making the current amplitude equal to the previous amplitude multiplied by some fractional scaling factor 'DR'. Consequently, if a '1' is detected as exiting the thresholding process for the current and previous 'C' blocks the previous B+C filter amplitudes for that particular frequency band have to be adjusted. For filter amplitudes in blocks (b-B-C+1) to (b-C-1) the amplitudes may be adjusted by a linear amount, for example  $\Delta A$ , where

$$\Delta A_{\frac{n}{NT}} = \left(1 - \text{Amp}_{b-B-C, \frac{n}{NT}}\right) / B \quad (18)$$

$\text{Amp}_b$  = current filter amplitude.

Then

$$\left\{ \text{Amp}_{b-B-C-k, \frac{n}{NT}} \right\}_{k=1}^{B-1} = \left\{ \text{Amp}_{b-B-C, \frac{n}{NT}} + k \cdot \Delta A_{\frac{n}{NT}} \right\}_{k=1}^{B-1} \quad (19)$$

(Note that any function rather than linear could be used but linear is simple to implement and works perfectly well for the majority of applications)

Filter amplitudes in blocks (b-C) to (b) have filter amplitudes for the nth frequency band equal to unity.

It can be seen therefore, that the depth of the data delay block 11 as shown in Figure 3(a) has to be at least (B+C).N in order to be able to go back and adjust the ideal filter amplitudes of the previous (B+C) blocks on the basis of the current blocks analysis. In order to be able to perform the analysis and adjust the ideal filter amplitudes



of the previous (B+C) blocks it is necessary to store both the 1's and 0's vector and filter amplitudes of the previous (B+C) blocks in stores 18 and 26 as shown in Figure 3(b). The same process is applied to each of the  $\left(\frac{N}{2}\right)+1$  unique filter frequency bands for every block.

It has been found useful to make the parameters of Attack rate (B), Decay rate (DR), min signal duration (C) and maximum attenuation (MA) operator controls so that a user can adjust to suit the audio material being processed. This is illustrated by user controls 20, 24 and 25, 23 and 22 respectively in Figure 3(b).

Figure 7 illustrates how the filter/frequency band amplitudes are adjusted from the ideal (process 19 in Figure 3(b)) and Figure 8 illustrates the process in full detail by means of a flow chart.

Note also that in Figure 7 the parameters B, C and MA are constants whereas the nth filter can have its own decay rate. The purpose of this is that, for some audio material, it has been found that the higher frequency bands should have a different decay rate to the lower frequency bands. In the majority of cases if two decay rates are specified, i.e.  $DR_{n=0}$  and  $DR_{n=\frac{N}{2}}$  (as illustrated by 24 and 25 in Figure 3b) a linear increment  $\Delta DR$  between them is sufficient. With reference to Figure 3 the computational

element 21 computes the following equations given the upper and lower decay rate parameters:

$$\Delta DR = \left( DR_{n=\frac{N}{2}} - DR_{n=0} \right) / \left( \frac{N}{2} \right) \quad (20)$$

and

$$\{DR_n\}_{n=0}^{\frac{N}{2}} = \{DR_{n=0} + n \cdot \Delta DR\}_{n=0}^{\frac{N}{2}} \quad (21)$$

The sequence  $\{DR_n\}_{n=0}^{\frac{N}{2}}$  is then output by 21 to the adjust ideal filter levels 19 and used as illustrated in Figure 8 in the computation of the modified filter levels.

The output in the b-th block of the 'adjust ideal filter levels' process 19 is the modified sampled frequency response  $\left\{ Amp_{b-B-C+1, \frac{n}{NT}} \right\}_{n=0}^{\frac{N}{2}}$  for the (b-B-C+1)-th block as illustrated in Figure 8. Note that, because of the need to go back and adjust the filter amplitudes of the previous (B+C-1) blocks in this way, it necessarily introduces an overall process or pipeline delay of the same amount.

Given the adjusted ideal filter level sequence  $\left\{ Amp_{b-B-C+1, \frac{n}{NT}} \right\}_{n=0}^{\frac{N}{2}}$  the process of obtaining the adaptive filter weights consists of firstly inverse transforming (27 in Figure 3b) the sequence to give the ideal impulse response of the filter. That is, for the case of the IDFT/IFFT and the (b-B-C+1)-th batch

$$\{h_{kT, (b-B-C+1)}\}_{k=0}^{N-1} = \left\{ \frac{1}{N} \cdot \sum_{n=0}^{N-1} Amp_{(b-B-C+1), \frac{n}{NT}} \cdot e^{-j \frac{2\pi nk}{N}} \right\}_{k=0}^{N-1} \quad (22)$$

where

$$\left\{ Amp_{(b-B-C+1), \frac{n}{NT}} \right\}_{n=1}^{\frac{N}{2}-1} = \left\{ Amp_{(b-B-C+1), \frac{n}{NT}} \right\}_{n=1}^{\frac{N}{2}-1} \quad (23)$$

The number of outputs required from the filter process is

$$N(1-\alpha) \text{ filtered output samples/batch} \quad (24)$$

where  $N$  is the number of points

$\alpha$  is the overlap ratio e.g. 0.5 if 50%

Therefore the number of stages within the finite impulse response filter should ideally be the maximum possible such that when convolved with an  $N$  point data sequence it produces  $N.(1-\alpha)$  filtered outputs.

For the case of 50% overlap this is equivalent to  $(\frac{N}{2})+1$  stages implying  $(\frac{N}{2}+1)$  filter coefficients. In this case, the filter coefficients are obtained from the following

$$\left\{ b_{kT + \frac{NT}{4}, (b-B-C+1)} \right\}_{k=0}^{\frac{N}{4}} = \left\{ h_{kT, (b-B-C+1)} \cdot w(2)_{kT + \frac{NT}{4}} \right\}_{k=0}^{\frac{N}{4}} \quad (25)$$

and

$$\left\{ b_{kT, (b-B-C+1)} \right\}_{k=0}^{\frac{N}{4}-1} = \left\{ b_{\frac{NT}{2} - kT, (b-B-C+1)} \right\}_{k=0}^{\frac{N}{4}-1} \quad (26)$$

and are computed by stage 29 in Figure 3b. Computing the filter coefficients as per equations (25) and (26) ensures that the coefficients are real-valued and symmetric which is a necessary and sufficient condition to ensure linear phase.

The pre-computed set of weights (stored in window function store 28 (Figure 3b))  $\{w(2)_{kT}\}_{k=0}^{\frac{N}{2}}$  would be typically -40 dB Dolph-Chebyshev (or similar).

The filter coefficient sequence  $\{b_{kT, (b-B-C+1)}\}_{k=0}^{\frac{N}{2}}$  is then convolved with the delayed original input data previously held in delay store 11 to produce the filtered  $\frac{N}{2}$  output sequence by the finite impulse response filter 30 as follows:

$$\{y_{kT+(b-B-C+1)\tau}\}_{k=0}^{\frac{N}{2}-1} = \left\{ \sum_{m=0}^{\frac{N}{2}} x_{(k+m)T+(b-B-C+1)\tau} \cdot b_{mT, (b-B-C+1)} \right\}_{k=0}^{\frac{N}{2}-1} \quad (27)$$

Notice that in equation (27) the designed filter coefficients are only valid for the  $(b-B-C+1)$ -th batch of data as they are completely re-computed for each subsequent batch. The main advantage of this technique is that the filter applied to the data has no a priori assumptions made about it but instead is designed strictly on the basis of the real-time analysis of a particular data batch and then only applied to that data. For the next batch the whole process is repeated.

The processed block of noise suppressed data is then re-assembled into a continuous digital data stream by a further double buffered memory 31 which stores one set whilst outputting the previous set.

Finally, two alternative output stages are provided, selectable by an output signal select switch 32. A digital audio signal output 38 is supplied direct with data from the double buffered memory 31 via a digital audio interface 37. Alternatively, data from store 31 is passed via switch

32 to a digital-to-analogue converter 33 whose output is low pass filtered at 34 in order to reconstruct the analogue signal before being passed to the analogue audio signal output 36 via a buffer amplifier 35.

It will be appreciated that whilst this exemplary embodiment of the invention has been described in terms of a time domain convolution, an alternative equally valid approach would be to perform a frequency domain multiplication, i.e. Fast Convolution, yielding a potentially more computationally efficient implementation.

Similarly, it will be appreciated that this exemplary embodiment of the invention has been described in terms of real-valued samples being taken of the original signal and where some appropriate advantage has been taken of this fact to reduce the computational requirements, i.e. halve the data arrays in certain circumstances. People skilled in the art will of course recognise that the method can be modified for the case of complex-valued samples e.g. Phase and Quadrature sampled baseband outputs of a communications receiver, or further computational advantage may be gained by performing multi-channels or multi-blocks simultaneously. For example, it is well known that it is possible to perform two simultaneous real-valued FFT's/DFT's with one complex-FFT. In this way the real-time computational requirements can be considerably reduced, or the size of the

data batch (implying resolution) can be increased for a given computational capability, or the block overlap ratio increased.

The invention has been particularly described in relation to audio noise reduction, as it is eminently suitable for use in the re-mastering of analogue recordings onto digital format or cleaning up poor quality recordings or restoring film sound tracks. However, it will be appreciated that the method and apparatus of the invention have much wider application in communications systems and in noise reduction in signals representing images, such as those found in Tomography, old movies etc, as well as signal detection systems more generally which require an accurate noise estimate to be established for subsequent use as a basis for a detection decision, adaptive actions or adaptive component signal extraction or suppression.

CLAIMS

1. Signal processing apparatus for use in discriminating between the component random signals, having different statistical distributions, in a composite signal consisting of a summation of two or more such random signals, comprising means for acquiring from a sequence of digital samples of said signal successive sets each containing a predetermined number of samples, and processing means for processing each successive set in turn, said processing means comprising calculating means for weighting and transforming the set of samples into a discrimination space and for computing a representation of the magnitude of the signal representation within the discrimination space, ordering means for sorting the discrete values or elements of the signal magnitude representation within the discrimination space to yield the data arranged in rank order, discriminating means for identifying a discrimination point or points within said rank ordered set, and thresholding means for dividing said rank ordered set in accordance with said discrimination point or points to form two or more sub-sets, the sub-sets substantially representing said component random signals of said composite signal, and means for identifying in and extracting from the original set of digital signal samples or the transformed set before rank ordering those elements which correspond to at least one of said component signals.

2. Apparatus according to Claim 1, wherein the discriminating means is arranged to identify a discrimination point or points within said rank ordered set at which there is a significant change in gradient.

3. Apparatus according to Claim 2, wherein the ordering means is arranged to sort the data set into ascending order of magnitude, and the discriminating means comprises means for calculating for each element in the sorted set an amplitude-independent discrimination sequence:-

$$g(d) = \frac{d \cdot P_{(d)}}{\sum_{r=0}^d P_{(r)}} \quad : d = 0, 1, \dots, N-1$$

wherein  $d$  is the index number of the magnitude data element in the rank ordered sequence,  $P$  is the magnitude and  $\{P_r\}_{r=0}^{N-1}$  is the rank ordered data sequence, and means for identifying the sub-set  $\{P_r\}_{r=0}^d$  representing those elements substantially from one of said component signals, such that, in the process of computing the said discrimination sequence  $g(d)$ , a value of  $d$  can be determined such that  $g(d)$  is less than or equal to a predetermined value and  $g(d+1)$  is greater than said predetermined value.

4. Apparatus according to Claim 3, wherein said predetermined value is in the range 1.9 to 2.4 for a magnitude representation.



5. Apparatus according to Claim 2, 3 or 4, which uses the magnitude  $P_{r..d}$  of the element at the discrimination point  $d$  as a basis of a thresholding process.

6. Apparatus according to any of Claims 2 to 5, comprising adjustment means arranged to adjust a discrimination point adaptively according to a statistical confidence factor derived from the angular difference or equivalent thereof in gradients of adjacent sub-sets.

7. Apparatus according to any preceding claim, wherein the discrimination space is the frequency domain and the calculating means is arranged to perform a Discrete or Fast Fourier Transform on said set, whereby said representation is the magnitude or power spectrum.

8. Apparatus according to any preceding claim, comprising means for suppressing those elements which correspond to a selected one of the sub-sets, and means for reconstructing the original composite signal with one, or at least one, of the component random signals substantially suppressed.

9. Apparatus for increasing the signal-to-noise ratio of an analogue signal containing additive random noise, comprising means for receiving a stream of digital samples of said signal and signal processing apparatus according to Claim 8, the suppressing means being arranged to suppress those elements which correspond to the sub-set having the

lower gradient (for ascending order ranking) substantially representing the random noise component of said analogue signal.

10. Apparatus according to Claim 9, comprising means for the real-time design of an adaptive filter for suppressing the data elements within a selected sub-set representing an unwanted component signal of the composite signal and means for applying the designed filter only to the original set of data, appropriately delayed, on which the analysis was conducted.

11. Apparatus according to Claim 10, comprising means for temporally monitoring the contents of each frequency band in the wanted signal sub-set and for suppressing any element which is not present within the same frequency band for a predetermined number C of successive sets of data.

12. Apparatus according to Claim 11, wherein means are provided for selectively varying the value of C.

13. Apparatus according to any of Claims 10 to 12, comprising means for monitoring the output of each frequency band exiting the thresholding process and for detecting a transition from noise to signal over successive sets of data, and means for controlling the adaptive filter amplitude for the said frequency band such that the filter amplitude of that frequency band in each of the preceding B sets of

data before the detected transition is increased by a small increment such that the filter amplitude has reached unity value by the noise to signal transition point.

14. Apparatus according to Claim 13, wherein means are provided for selectively varying the increment size and the value of B.

15. Apparatus according to any of Claims 10 to 14, comprising means for monitoring the output of each frequency band exiting the thresholding process and for detecting a transition from signal to noise over successive sets of data, and means for reducing the adaptive filter amplitude for each of D data sets subsequent to the signal to noise transition from unity to some predetermined maximum suppression level.

16. Apparatus according to Claim 15, wherein means are provided for selectively varying the filter amplitude decay rate and function.

17. Apparatus according to Claim 15 or 16, wherein means are provided for selectively varying the maximum suppression level between 1 and 0.

18. Apparatus according to Claim 15, 16 or 17, wherein means are provided for selectively varying the value of D.

19. Apparatus according to any of Claims 8 to 18, comprising means for converting said analogue signal into said stream of digital samples.

20. Apparatus according to any of Claims 1 to 6, comprising means for selecting those samples in the transformed set which correspond to a selected one of the sub-sets, and means for further processing of said selected samples.

21. Apparatus for detecting a desired signal in a composite signal comprising the desired signal and random noise, means for receiving a stream of digital samples of said composite signal, and signal processing apparatus according to Claim 20 for processing said samples, wherein said further processing means is adopted to carry out a detection decision based on an ordered statistics estimate of the noise component of said composite signal.

22. Apparatus according to any preceding claim, wherein digital to analogue converter means are provided for converting the processed digital signal representation back into an analogue form.

23. Apparatus according to any preceding claim, wherein the acquiring means is arranged such that the successive sets of samples overlap.

24. Apparatus according to Claim 23, wherein the amount of overlap is at least 50%.

25. A signal processing method for use in discriminating between the component random signals, having

different statistical distributions, in a composite signal consisting of a summation of two or more such random signals, comprising:

(a) acquiring from a sequence of digital samples of said signal a set containing a predetermined number of samples;

(b) weighting and transforming the set of said signal samples into a discrimination space;

(c) computing a representation of the magnitude of the signal representation within the discrimination space;

(d) sorting discrete values of the signal magnitude representation within the discrimination space to yield the data arranged in rank order;

(e) identifying a discrimination point or points within said rank ordered set;

(f) dividing said rank ordered set in accordance with said discrimination point or points to form two or more sub-sets, the sub-sets substantially representing said component random signals of said composite signal;

(g) identifying in or extracting from the original set of digital signal samples or the transformed set before rank ordering those elements which correspond to at least one of said component signals; and

(h) repeating steps (a) to (g) with successive sets of samples.

26. A method according to Claim 25, wherein the or each discrimination point is a point at which there is a significant change in gradient of the rank ordered set.

27. A method according to Claim 26, comprising sorting the data set into ascending order of magnitude and calculating for each element in the sorted set an amplitude-independent discrimination sequence:-

$$g(d) = \frac{d \cdot P_{(d)}}{\sum_{r=0}^d P_{(r)}} \quad : d = 0, 1, \dots, N-1$$

wherein  $d$  is the index number of the magnitude data element in the rank ordered sequence,  $P$  is the magnitude and  $\{P_r\}_{r=0}^{N-1}$  is the rank ordered data sequence, and identifying the sub-set  $\{P_r\}_{r=0}^d$  representing those elements substantially from one of said component signals, such that, in the process of computing the said discrimination sequence  $g(d)$ , a value of  $d$  can be determined such that  $g(d)$  is less than or equal to a predetermined value and  $g(d+1)$  is greater than said predetermined value.

28. A method according to Claim 27, wherein said predetermined value is in the range 1.9 to 2.4 for a magnitude representation.

29. A method according to Claim 26, 27 or 28, comprising using the magnitude  $P_{r=d}$  of the element of the discriminating point  $d$  as a basis of a thresholding process.

30. A method according to any of Claims 26 to 29, wherein the discrimination space is the frequency domain and the transforming step is performed by means of a Discrete or Fast Fourier Transform, whereby said power representation is the magnitude or power spectrum.

31. A method according to any of Claims 26 to 30, comprising adaptively adjusting the discrimination point according to a statistical confidence factor derived from the angular difference or equivalent thereof in gradients of adjacent sub-sets.

32. A method according to any of Claims 26 to 31, comprising suppressing those elements which correspond to a selected one of the sub-sets, and reconstructing the original composite signal with one, or at least one, of the component random signals substantially suppressed.

33. A method for increasing the signal-to-noise ratio of an analogue signal containing additive random noise, comprising receiving a stream of digital samples of said signal, and processing said samples according to the method specified in Claim 32, wherein those elements which correspond to the sub-set having the lower gradient (for ascending order ranking) substantially representing the random noise component of said analogue signal.

34. A method according to Claim 33, comprising designing in real-time an adaptive filter for suppressing the data elements within a selected sub-set representing an

unwanted component signal of the composite signal and applying the designed filter only to the original set of data, appropriately delayed, on which the analysis was conducted.

35. A method according to Claim 34, comprising temporally monitoring the contents of each frequency band of each sub-set of processed samples corresponding to the desired signal, and suppressing any element in the sub-set which is not present within the same frequency band for a predetermined number of successive such sub-sets of data.

36. A method according to Claim 35, wherein said predetermined number is selectively variable.

37. A method according to Claim 34, 35 or 36, comprising monitoring the output of the thresholding process for each frequency band in each sub-set of processed samples and detecting in said frequency band a transition from noise to signal over successive such sub-sets of data, and adjusting the adaptive filter amplitudes for the frequency band such that the filter amplitude in the frequency band in each of the preceding B sets of data before the detected transition is increased by a small increment such that the filter amplitude has reached unity value by the noise to signal transition point.

38. A method according to Claim 37, wherein the increment size and the value of B are selectively variable.



39. A method according to any of Claims 34 to 38, comprising monitoring the output of the thresholding process for each frequency band in each sub-set of processed samples and detecting a transition from signal to noise over successive sets of data, and reducing the adaptive filter amplitude for each of D data sets subsequent to the signal to noise transition from unity to some predetermined maximum suppression level.

40. A method according to Claim 39, comprising selectively varying the filter amplitude decay rate and function.

41. A method according to Claim 39 or 40, comprising selectively varying the maximum suppression level between 1 and 0.

42. A method according to Claim 39, 40 or 41, comprising selectively varying the value of D.

43. A method according to any of Claims 33 to 38, comprising converting said analogue signal into said stream of digital samples.

44. A method according to any of Claims 25 to 43, wherein the successive sets of samples are selected so as to overlap.

45. A method according to Claim 44, wherein the amount of overlap is at least 50%.

46. A method according to any of Claims 25 to 31, comprising selecting those samples in the transformed set which correspond to a selected one of the sub-sets, and further processing said selected samples.

47. A method for detecting a desired signal in a composite signal comprising the desired signal and random noise, comprising receiving a stream of digital samples of said signal, and processing said samples according to the method specified in Claim 46, said further processing comprising carrying out a detection decision based on an ordered statistics estimate of the noise component of said composite signal.

48. An audio recording, when processed by a method according to any of Claims 33 to 45.

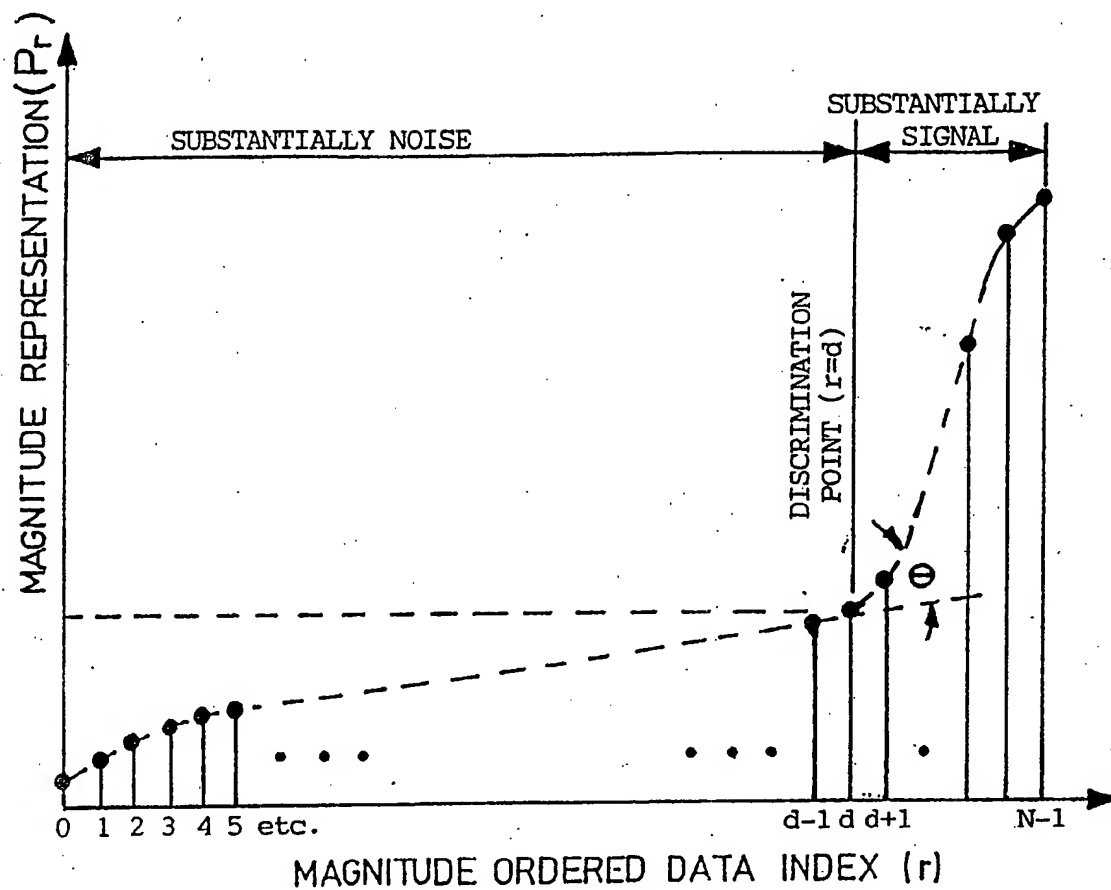


Figure 1

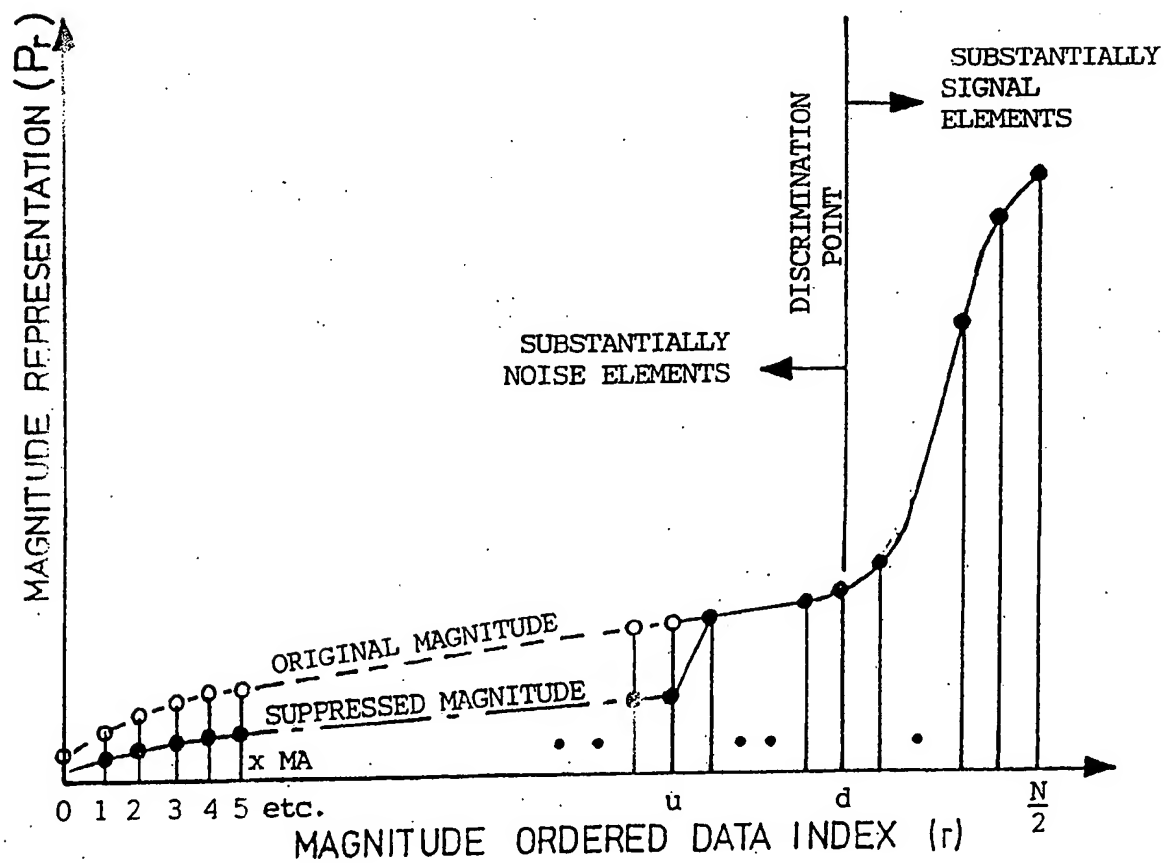


Figure 4

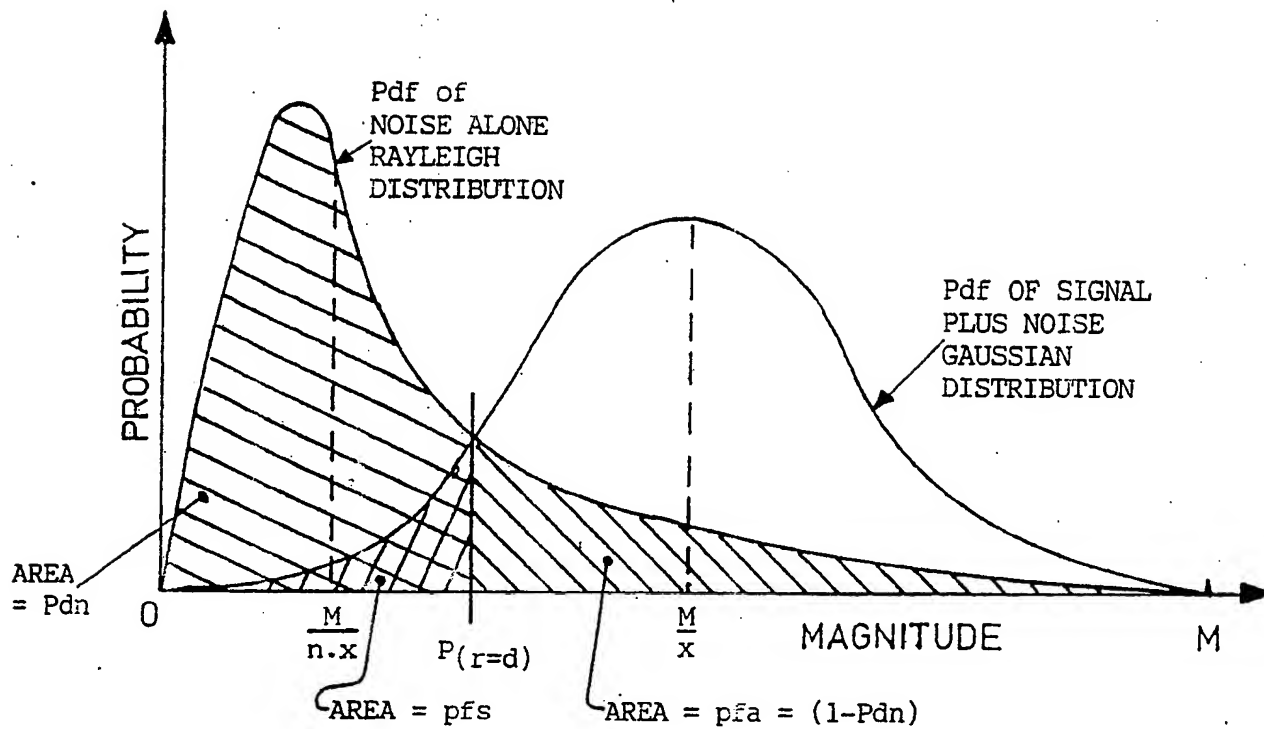


Figure 2a

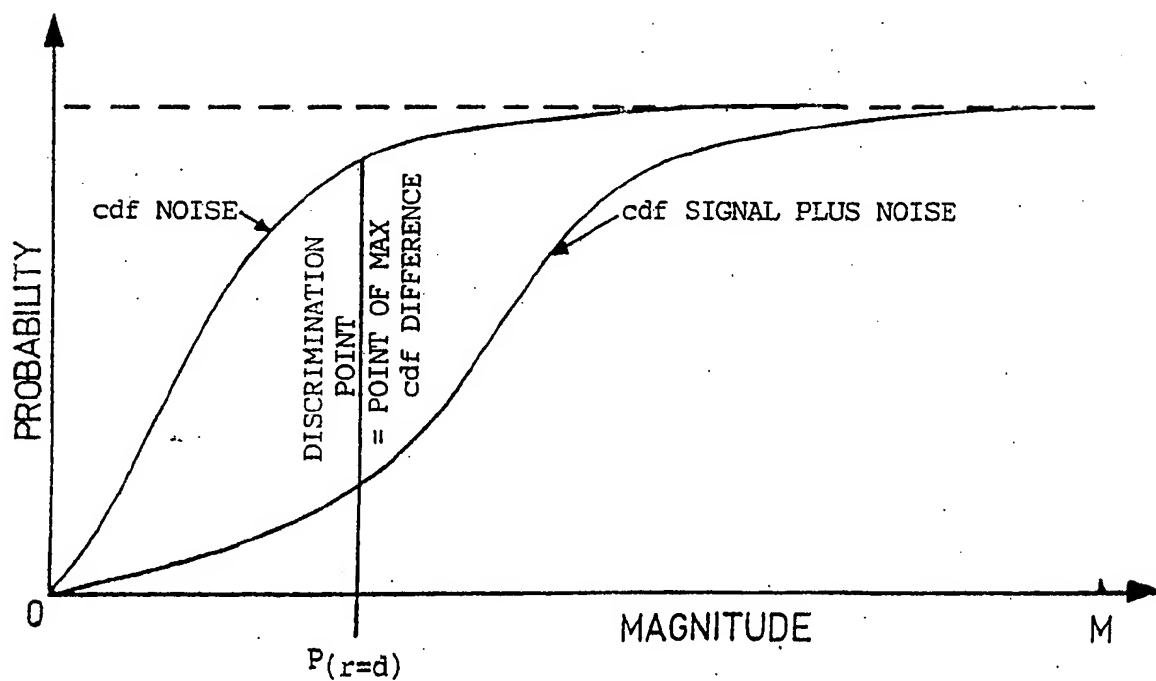
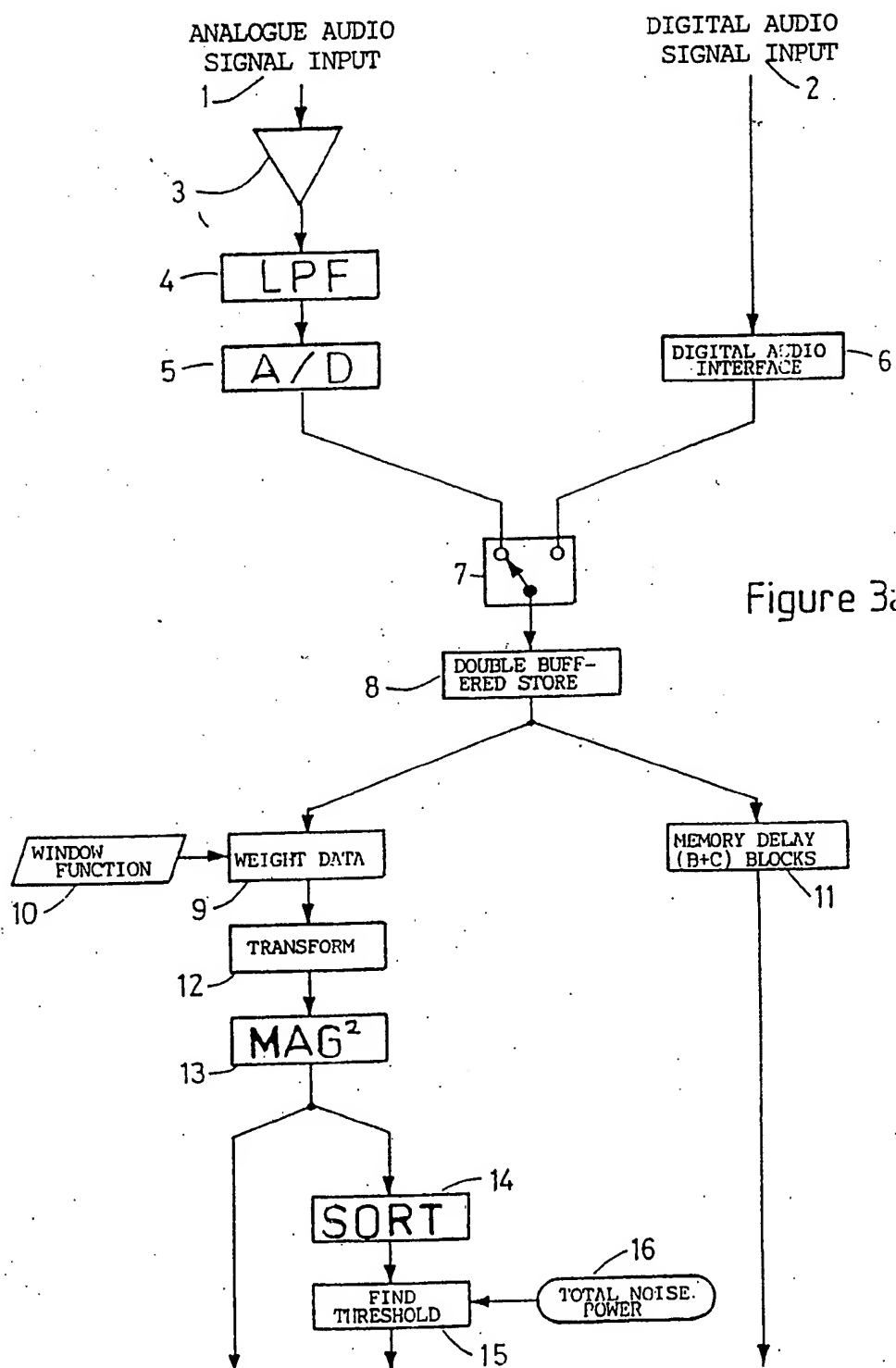
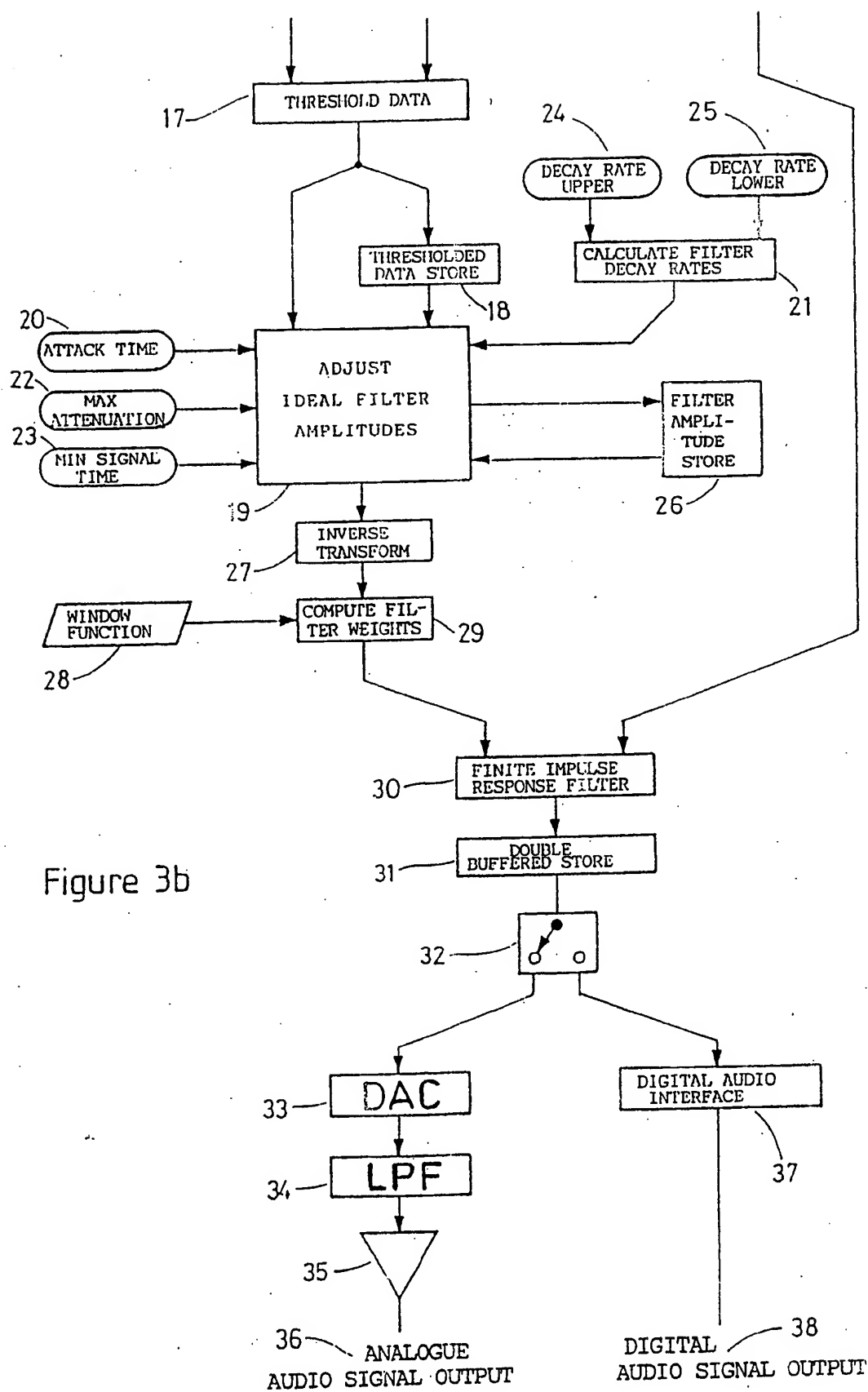


Figure 2b





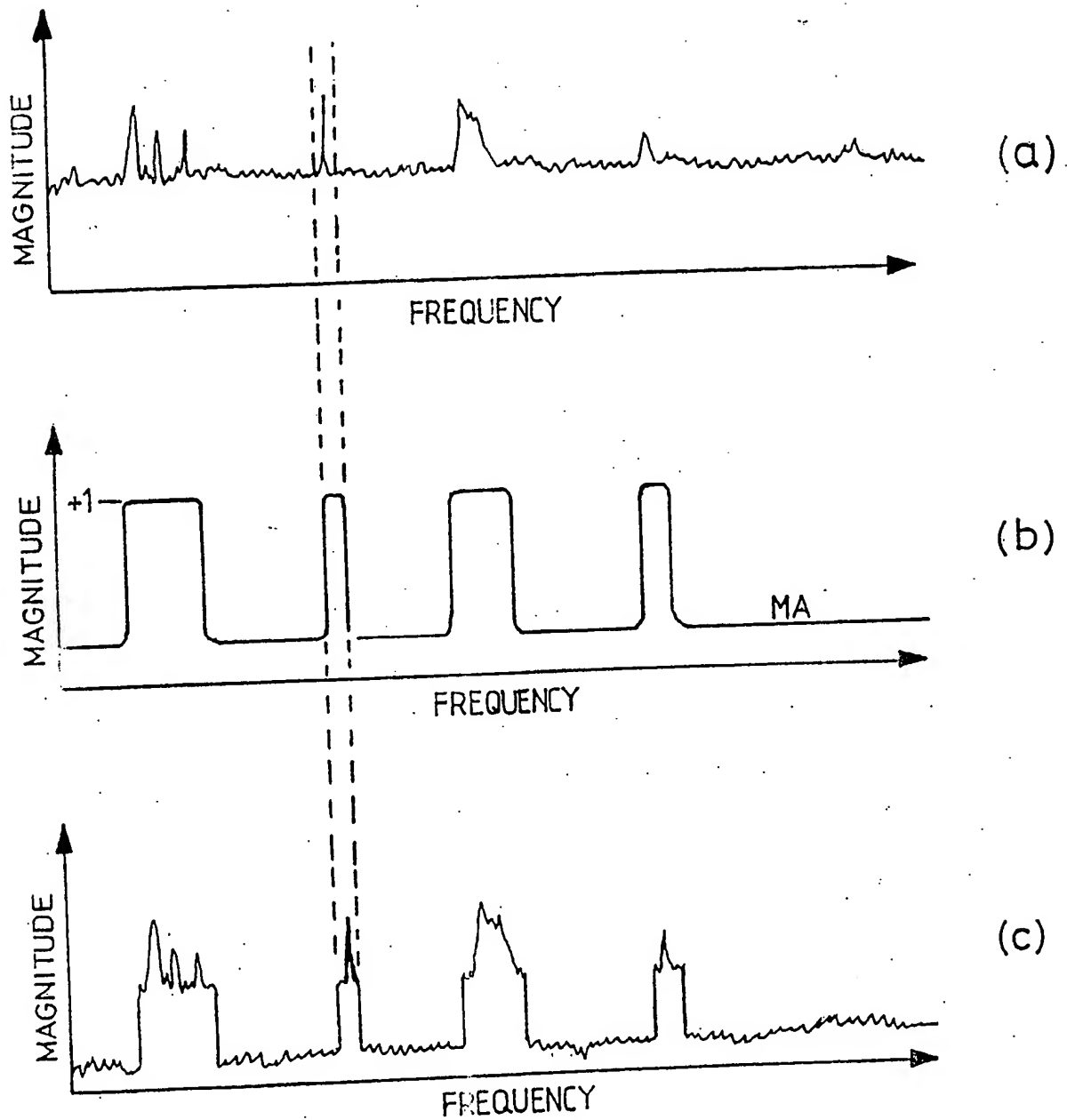


Figure 5

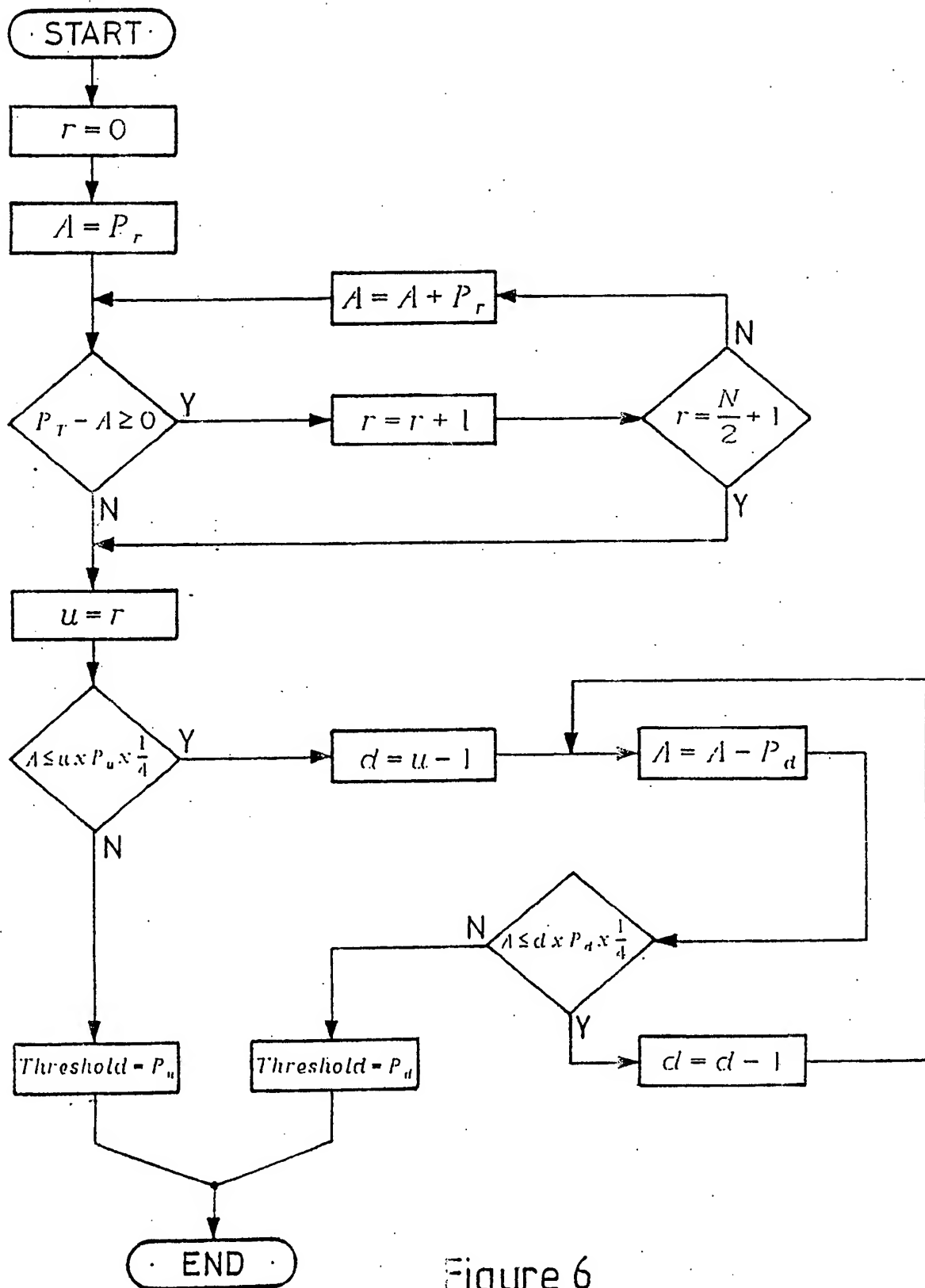


Figure 6



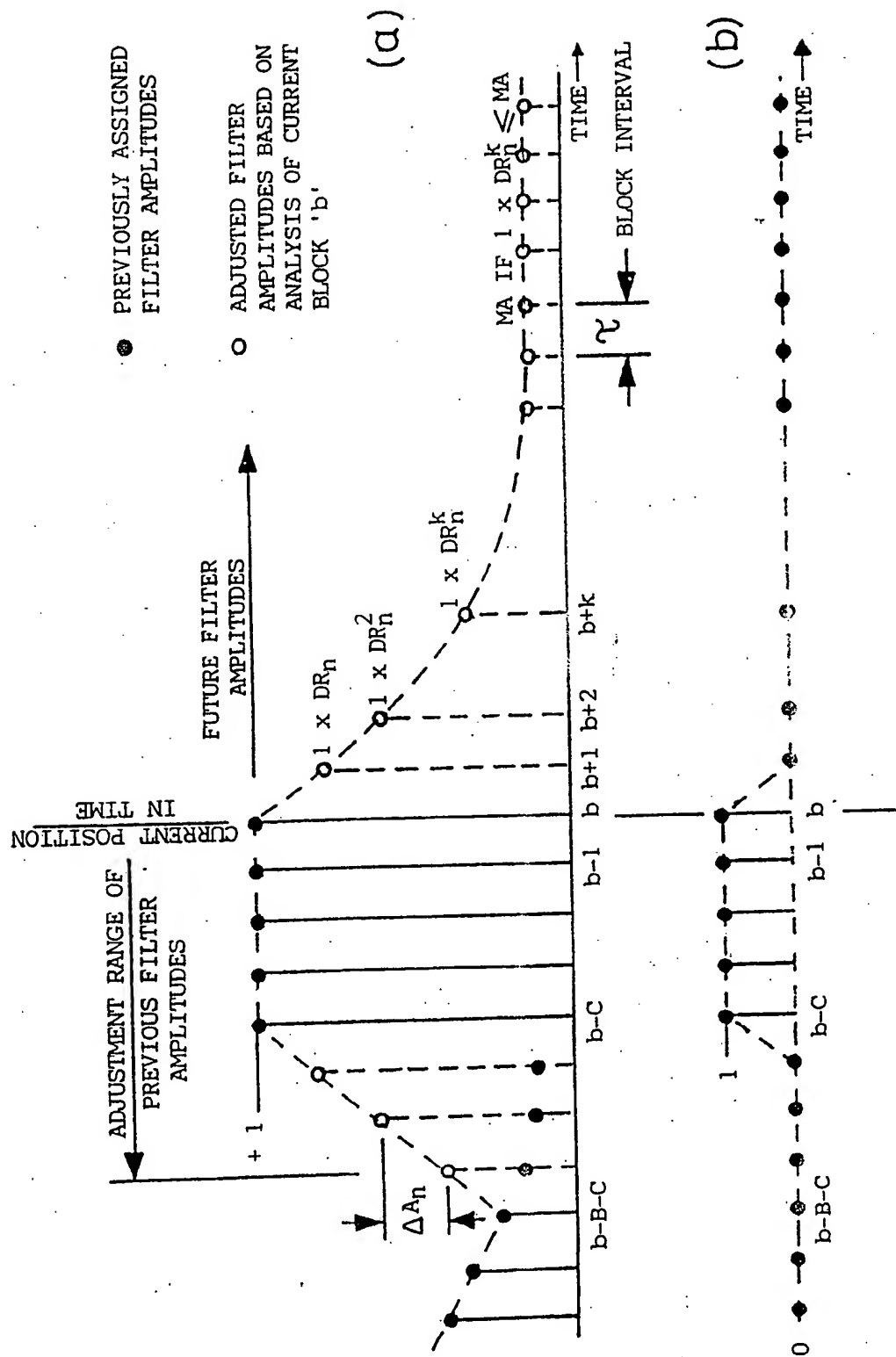
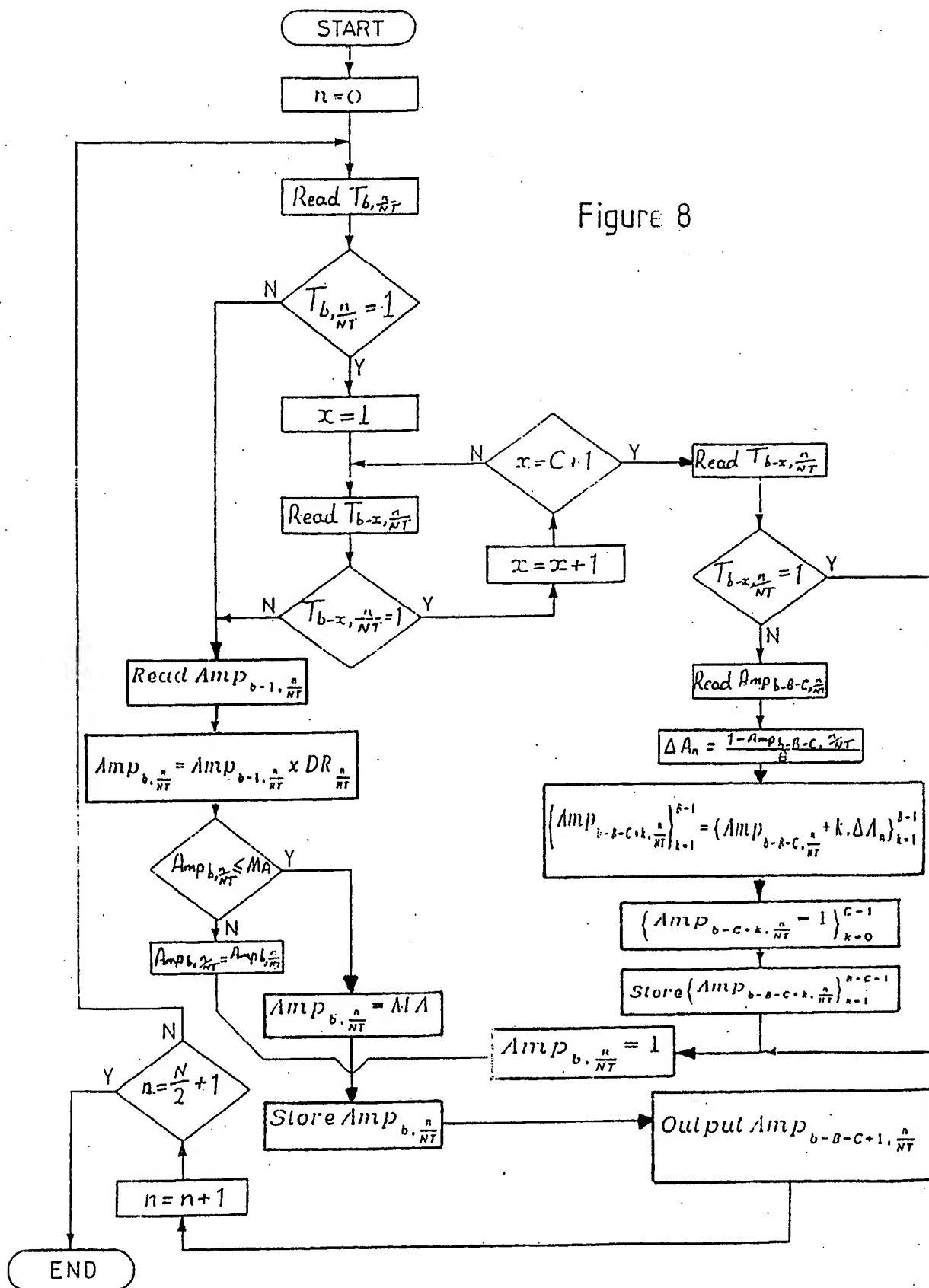


Figure 7

Figure 8



1

III. DOCUMENTS CONSIDERED TO BE RELEVANT (CONTINUED FROM THE SECOND SHEET)		
Category *	Citation of Document, with indication, where appropriate, of the relevant passages	Relevant to Claims No.
A	<p>INTERNATIONAL CONFERENCE ON ACOUSTICS, SPEECH AND SIGNAL PROCESSING vol. 2, 26 May 1989, GLASGOW, GB pages 1349 - 1352; BILGUTAY: 'Frequency diverse statistic filtering for clutter suppression' see page 1349, left column, line 39 - page 1350, left column, line 39; figure 1</p> <p>---</p>	1,7,25